Solid State Micro Technology
for Music

1979 DATA BOOK 1979

Solid State Micro Technology
2076B Walsh Avenue
Santa Clara, California 95050
Telephone: **(408) 248-0917**
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SECTION I
VOLTAGE CONTROLLED AMPLIFIERS

Solid State Micro Technology for Music

SSM 2000

DUAL LINEAR-ANTILOG VOLTAGE CONTROLLED AMPLIFIER

DESCRIPTION

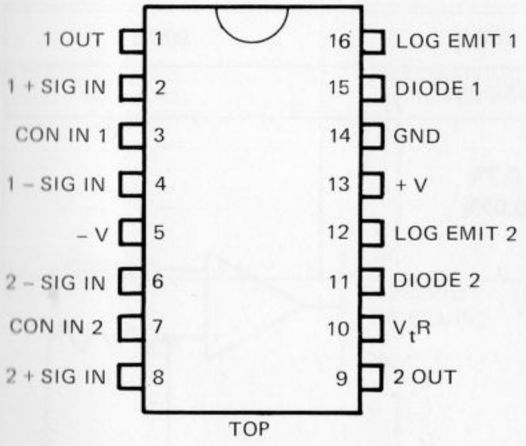
The SSM 2000 is a dual two-quadrant multiplier, each channel having separate control and differential signal inputs and a current output. In addition to linear amplitude control, on chip logging elements have been provided for producing an antilog control characteristic at the option of the designer. The device may be used in a wide variety of audio frequency applications including Voltage Controlled Amplifiers, AGC Circuits and as a Biquad Tuning Element. Both channels are temperature compensated and each channel has an 80 dB range.

FEATURES

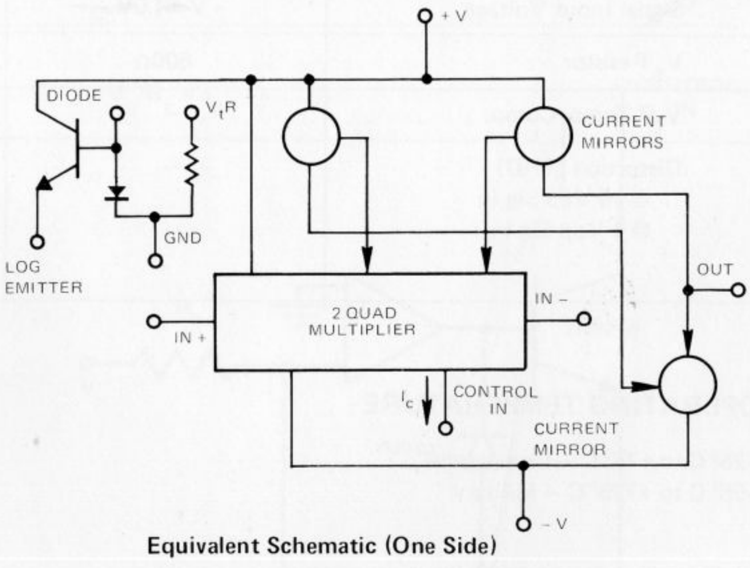
- Linear or Antilog Control Characteristics
- 80 db Control Range
- Excellent Control Accuracy (1% or better over the entire control range)
- Dual Design (completely independent selection of control characteristics)
- Internal Temperature Compensation
- Current Output
- Differential Signal Inputs
- 200 Nanoamp Input Bias Current
- Low Noise
- Low Distortion

APPLICATIONS

- Voltage Controlled Filters
- Two and Four Quadrant Multipliers
- Volume Controls
- Dividers
- AGC Circuits
- Voltage Controlled Current Sources and Sinks
- Compandors
- Voltage Controlled Quadrature Oscillators



Pin Diagram



Equivalent Schematic (One Side)

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SPECIFICATIONS

@25 @ 25°C, $V_s = \pm 12V$

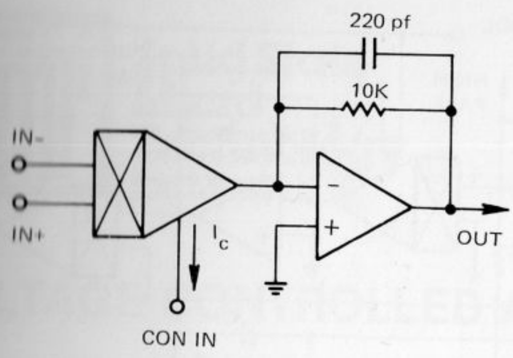
PARAMETER	MIN	TYP	MAX
Signal Input Bias Current	–	200nA	400nA
Noise Sig/Noise @ $10V_{pp}$ Sig in	–	84 db	–
gm gm Tempco	1/11kΩ	1/12kΩ 100 ppm/C°	1/13kΩ
Channel Matching a – Linear b – Antilog	 – –	 1% 2.5%	 2.5% 10%
Channel Separation @ 1 KHz	–	100db	–
Bandwidth I_c = lmA = 10 μA = 100nA	 – – –	 800kHz 100KHz 30kHz	 – – –
Supply Voltage	± 6V	± 10V	± 12V
Control Current	–	–	1 mA
AC Input Resistance	50 Meg	100 Meg	–
I offset/I control (untrimmed)	–	5%	12%
Control Rejection (trimmed)	–	50 db	–
Supply Current @ I control = 1 mA	–	10mA	12 mA
Signal Input Voltage	- V +4.0V	–	+ V -2.5V
V_t Resistor	600Ω	700Ω	900Ω
V_tR Temp. Comp.	–	+2000ppm/C	–
Distortion (THD) @ 10 Vpp Sig In @ 1 Vpp Sig In	–	 0.7% 0.07%	–

OPERATING TEMPERATURE

-25° C to +75° C – Commercial
-55° C to +125° C – Military

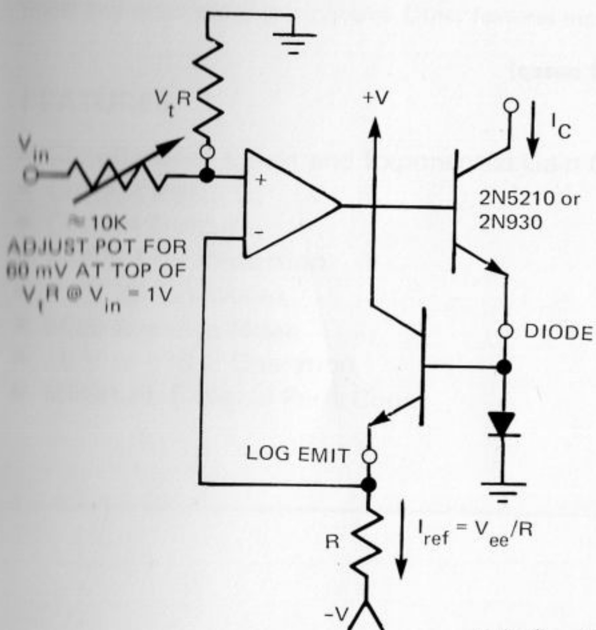
STORAGE TEMPERATURE

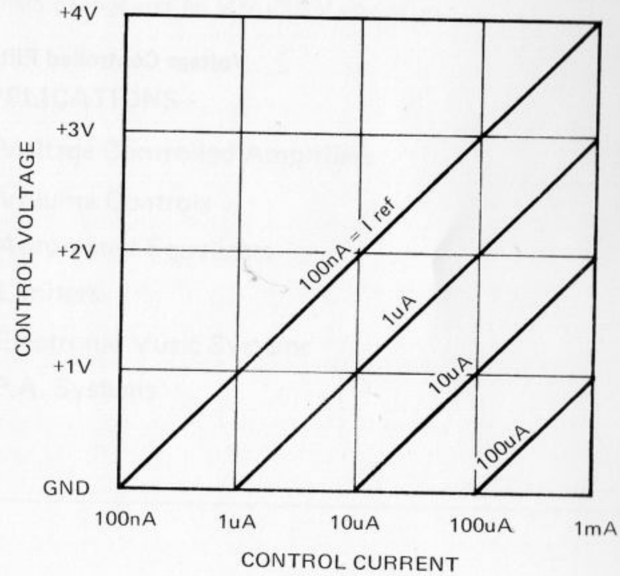
-55° C to +125° C



$$V_{out} = \frac{I_c\,(V+ - V-)\ 10k}{11.8V}$$

VCA





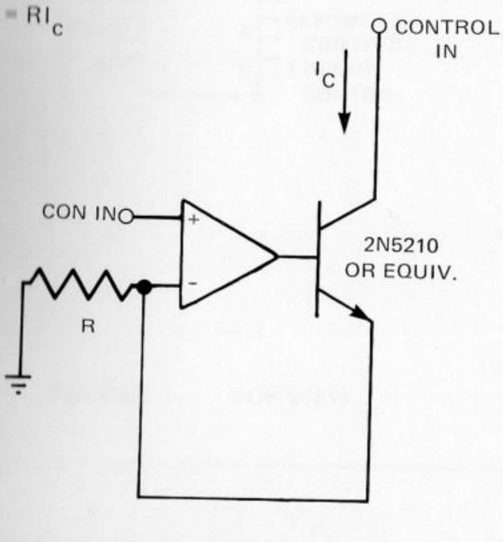
$$\ln\,(I_c/I_{ref}) = R_1\,V_{in}/V_T\,(R_1 + R_2)$$

$$V_T = \frac{KT}{q} = 26\ \text{m V @ } 25^\circ\text{C}$$

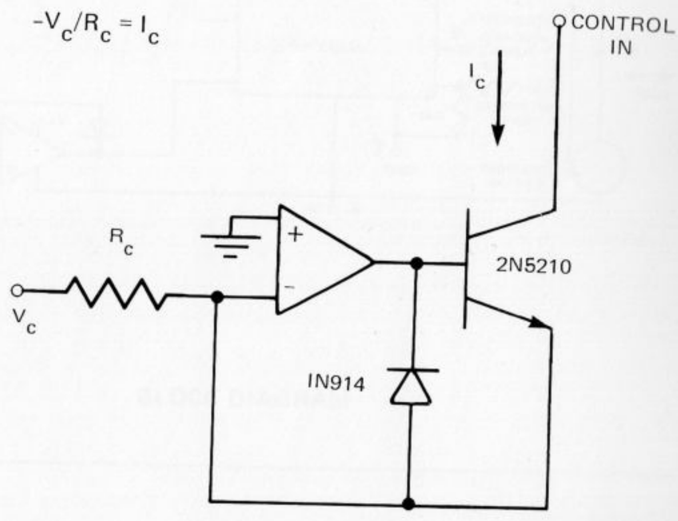
Antilog Control Circuit

Antilog Control Graph

$V_c = RI_c$



$-V_c/R_c = I_c$



Positive Linear Control Circuit

Negative Control Circuit

VCF Design Equations

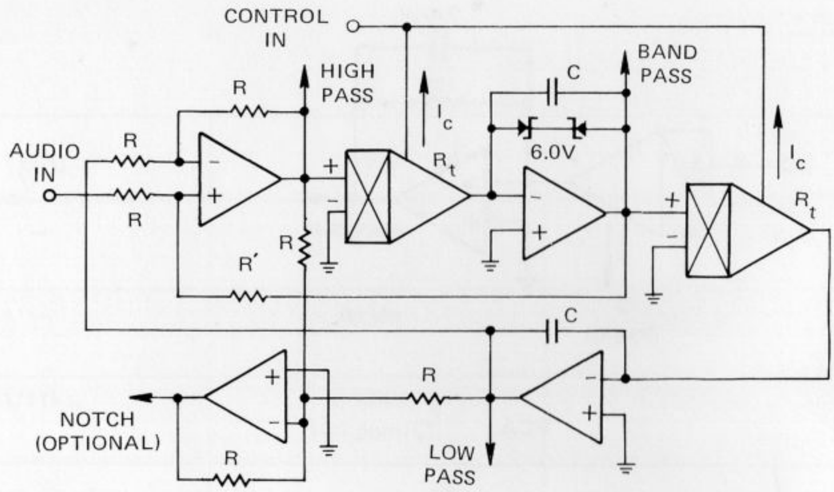
$R_t C w_c = 1$

$R_t = 21.2K$ @ $I_c = 500uA$

$R' = (2Q-1)R$

$A = 2-1/Q$

$R = 10K$



Voltage Controlled Filter (10,000 to 1 sweep)

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SSM 2010

VOLTAGE CONTROLLED AMPLIFIER

DESCRIPTION

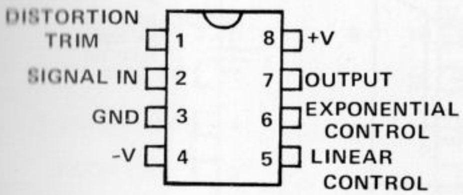
The SSM 2010 is a precision two quadrant multiplier designed for quality electronic music and P.A. systems. The device offers very low distortion and high signal/noise ratio, a minimum external parts count and a complete on-chip control circuit for simultaneous linear and exponential gain control. Other features include a wide dynamic range and ±5 V to ±18 V operation.

FEATURES

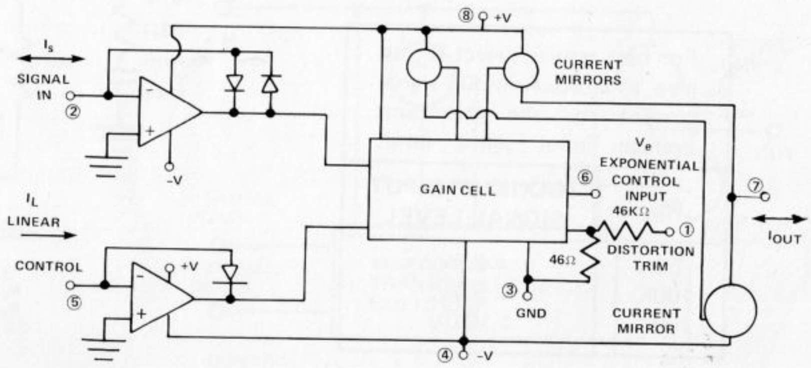
- Simultaneous Linear and Exponential Gain Control.
- Current Input
- Current Output
- 0.05% THD Distortion
- 0.2% IM Distortion
- 90db Signal-to-Noise
- ±5 V to ±18 V Operation
- Minimum External Parts Count

APPLICATIONS

- Voltage Controlled Amplifiers
- Volume Controls
- Automated Equalizers
- Limiters
- Electronic Music Systems
- P.A. Systems



PIN OUT TOP VIEW



BLOCK DIAGRAM

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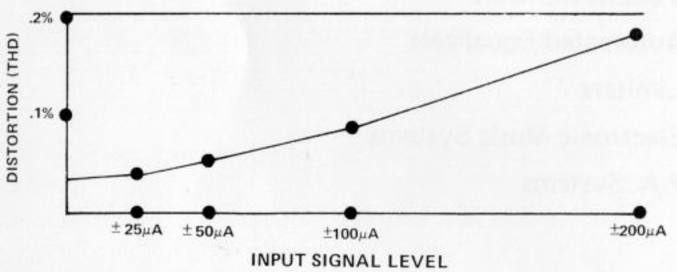
SPECIFICATIONS

$V_S = \pm 15$ V, $T_A = 25°$ C (unless otherwise specified).

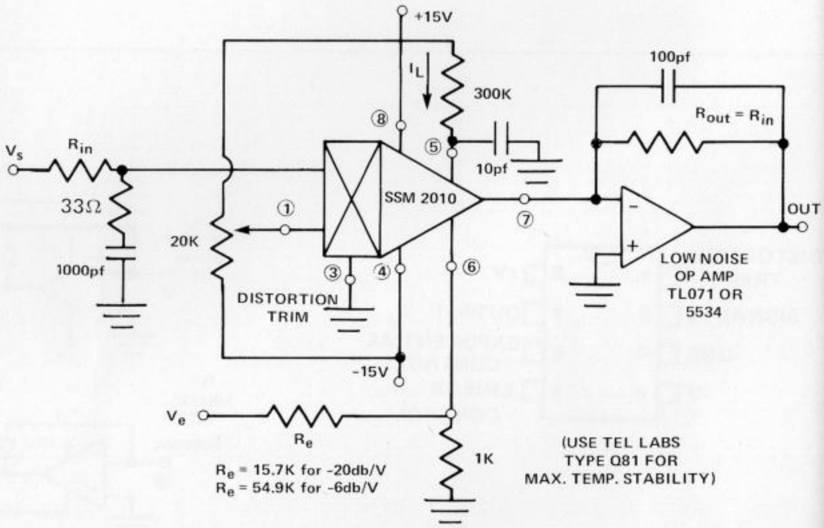
STORAGE TEMPERATURE -55 – +125°C

OPERATING TEMPERATURE 0 to 75°C

PARAMETER	MIN	TYP	MAX	CONDITIONS
I_{CC}	3.0mA	5.0mA	10mA	$V_S = \pm 18$ V
Output offset	-5μA		+5μA	$I_s = 0$ $V_e = 0$ $I_L = 50\mu A$
Gain	0.75	1	1.25	$I_s = \pm 100\mu A$ $V_e = 0, I_L = 50\mu A$
Peak Output	±200μA pp	±300μA pp		$I_s = \pm 300\mu A$ $V_e = 0, I_L = 50\mu A$
Output Leakage	-100nA		+100nA	$I_s = 0, I_L = 0$ $V_e = 0$
Expo Control Sensitivity		-6db/18mV		



For best results, select R_{in} to give a ± 50μA input signal current for the maximum average input signal level.	
R_{in}	MAXIMUM INPUT SIGNAL LEVEL
50K	± 2.5V
100K	± 5.0V
200K	± 10.0V



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SSM 2020

DUAL LINEAR-ANTILOG VOLTAGE CONTROLLED AMPLIFIER

DESCRIPTION

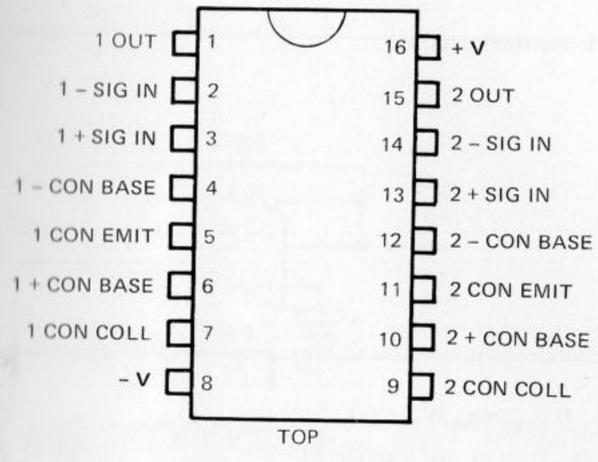
The SSM 2020 is a dual two quadrant multiplier designed to be used with op amps in a wide variety of precision audi-frequency applications including AGC circuits, Dividers and as a Biquad tuning element. Each channel has separate control and differential signal inputs and a current output. The device offers an exceptionally flexible control circuit for each channel which allows simultaneous linear and exponential voltage control of gain or either polarity of current control. Both channels are fully temperature compensated and have 86 dB signal-to-noise ratios at less than 0.1% distortion

FEATURES

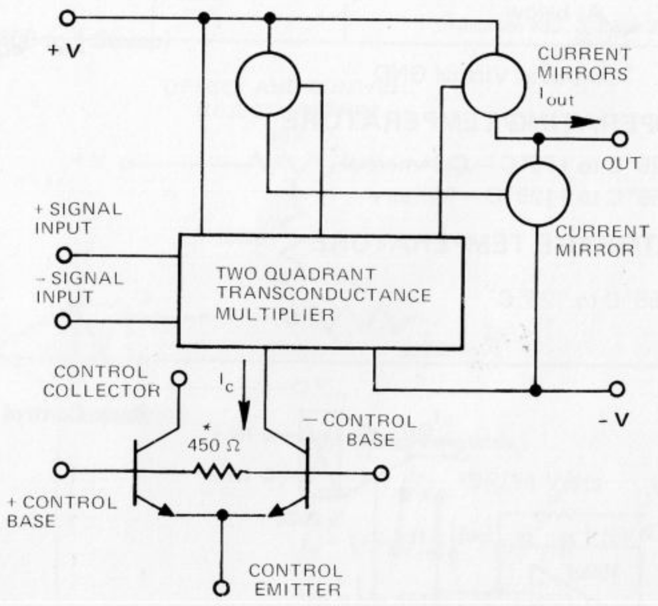
- Max Supplies ± 18V
- Dual Design (Independent Control Selection)
- 2% Channel Gain Matching
- 100 dB Control Range
- Simultaneous Linear and Exponential Gain Control
- Differential Signal Inputs
- Current Output
- 86db Signal-to-Noise
- 0.1% Distortion
- Fully Temperature Compensated

APPLICATIONS

- 2 and 4 Quadrant Multipliers
- Dividers
- AGC Circuits
- Voltage Controlled Filters
- Voltage Controlled Quadrature Oscillators
- Volume Controls
- Equalizers
- Compandors
- Antilog Amplifiers
- Voltage Controlled Current Sources



Pin Diagram



Equivalent Schematic (One Side)

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SPECIFICATIONS

$V_S = \pm 15V$, $I_{c1} = I_{c2} = 500\ \mu A$ and $T_A = 25°C$, unless otherwise specified.

PARAMETERS	MIN	TYP	MAX	CONDITIONS
Signal Input Bias I_b		500 nA	2.2 μA	$V_{ee} + 3V \leqslant V_+, V_- \leqslant V_{cc} - 3V$
Supply Voltage V_s	±6	±15	±18	
Supply Current I_s		6 mA	8 mA	$I_{c1} = I_{c2} = 1$ mA
Control Current			1 mA	
Transconductance gm	1/12kΩ	1/14 kΩ	1/16 kΩ	$I_{c1} = I_{c2} = 1$ mA
gm match		+2%	±5%	
gm Temco		100 ppm/°C		
Control Circuit V_{os}		1 mV	3 mV	
Output Offset I_o/I_c		±2%	±10%	$V_+ = V_- =$ GND (untrimmed)
Control Rejection		60 dB		$0 \leqslant I_c \leqslant 1$ mA (trimmed)
450 Ω Resistor	350 Ω	450 Ω	550 Ω	
450 Ω Temp Coef		+2000 ppm/°C		
Channel Separation		100 dB		F = 1 kHz
Bandwidth (3 dB)		1 MHz		$I_c = 1$ mA *
		300 kHz		$I_c = 10\ \mu A$
		30 kHz		$I_c = 100$ nA
Feedthrough:				
-Input to Output		90 dB		F = 1 kHz, $I_c = 0$
+ Input to Output		100 dB		F = 1 kHz, $I_c = 0$
Signal/Noise		86 dB		$V_s = 6\ V_{pp}$, $I_c = 1$ mA
Distortion (THD)				
VCA (Open Loop)		0.1%		$V_s = 6\ V_{pp}$, $I_c = 1$ mA
VCF (Closed Loop)		0.02%		$V_s = 6\ V_{pp}$, $I_c = 1$ mA
As below				

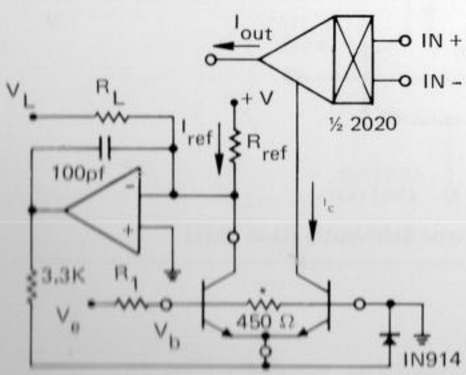
*Output at Virtual GND

OPERATING TEMPERATURE

-25° C to +75° C – Commercial
-55° C to +125° C – Military

STORAGE TEMPERATURE

-55° C to 125° C



Basic Control Circuit

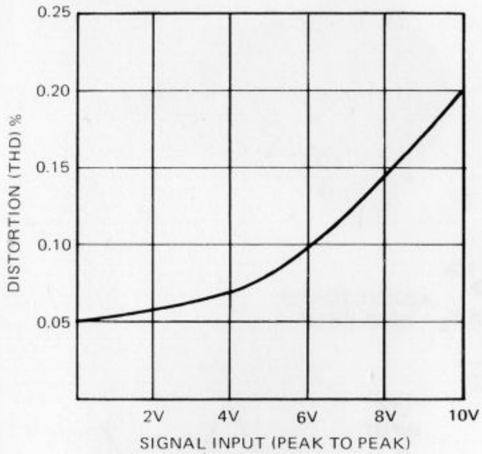
1) $I_{out} = g_m (V_+ - V_-)$

2) $I_{out} = \dfrac{I_c (V_+ - V_-)}{14 \text{ volts}}$

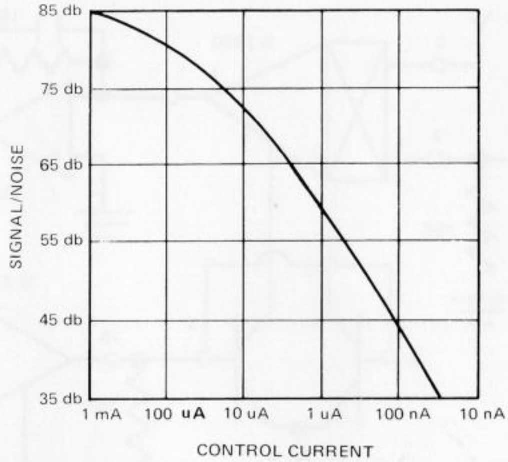
3) $I_c = e^{-V_b q/KT} (+V/R_{ref} + V_L/R_L)$

where $V_b = \dfrac{V_c\ 450\Omega}{R_1 + 450\ \Omega}$

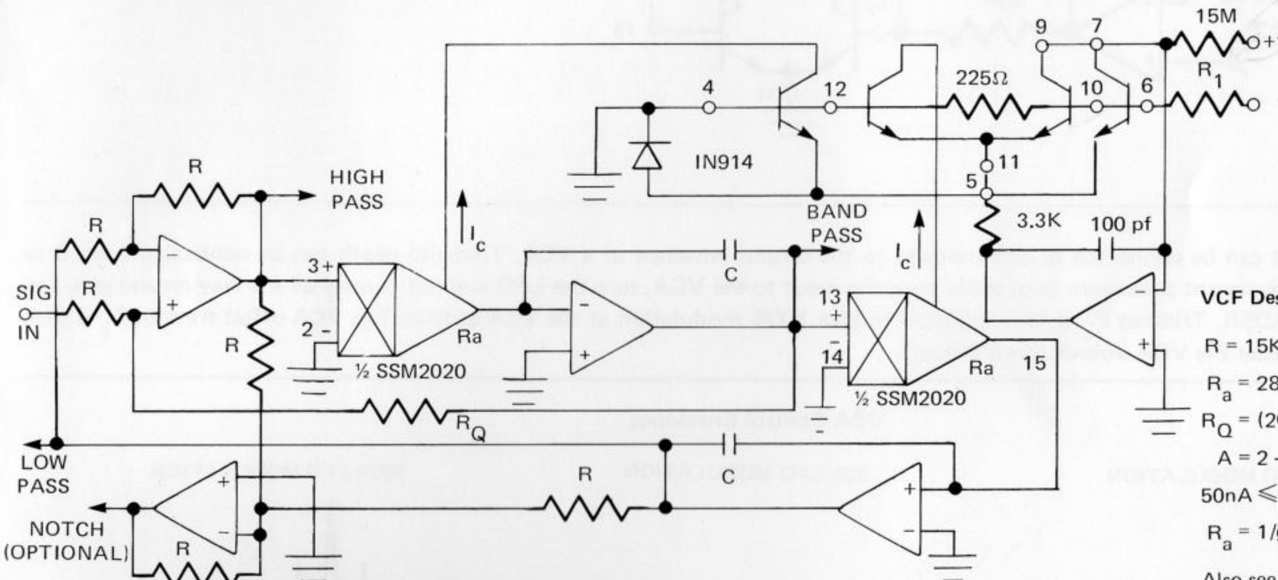
*NOTE: THE 450 Ω RESISTORS ARE INTERNAL TO THE I.C. AND COMPENSATE FOR THE T FACTOR IN THE EXPONENT.



Distortion vs Signal Input



Signal Noise vs Control Current (6 Vpp in)



Voltage Controlled Filter (10,000 to 1 Sweep)

VCF Design Equations

$R = 15K, \quad R_a C\omega_o = 1$

$R_a = 28K$ @ $I_c = 500$ uA

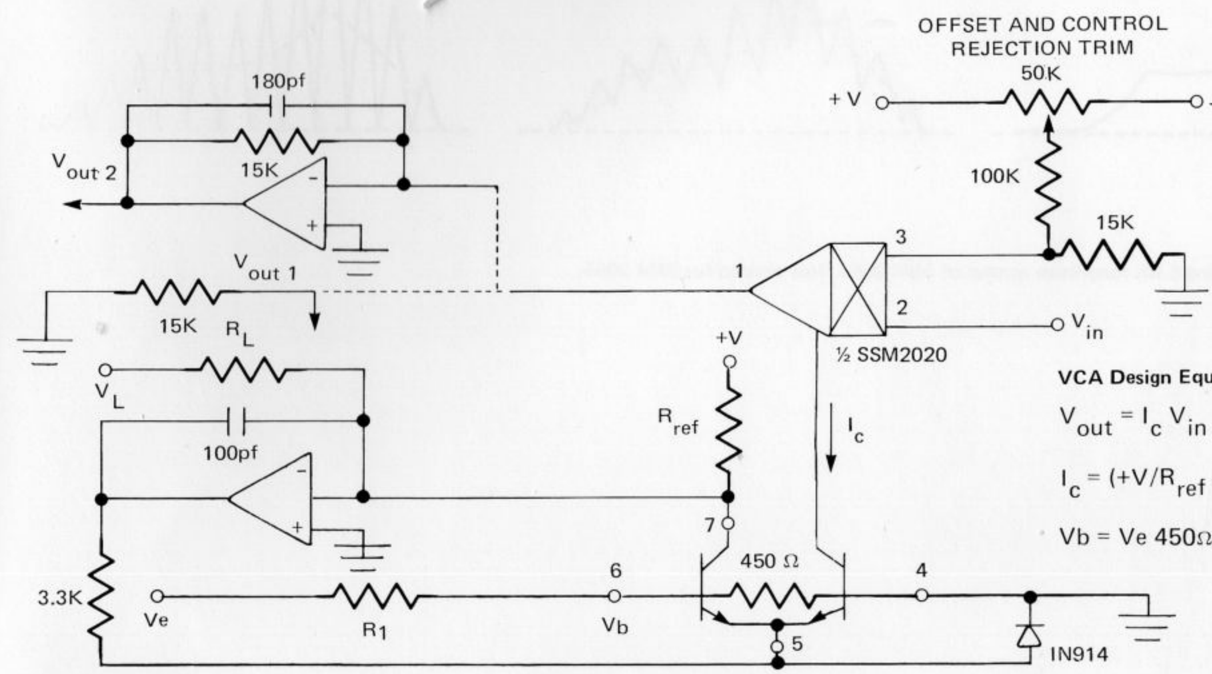
$R_Q = (2Q - 1)\ R$

$A = 2 - 1/Q$

$50nA \leqslant I_C \leqslant 500$ uA

$R_a = 1/gm = 14V/I_c$

Also see equ. 3, Page 2



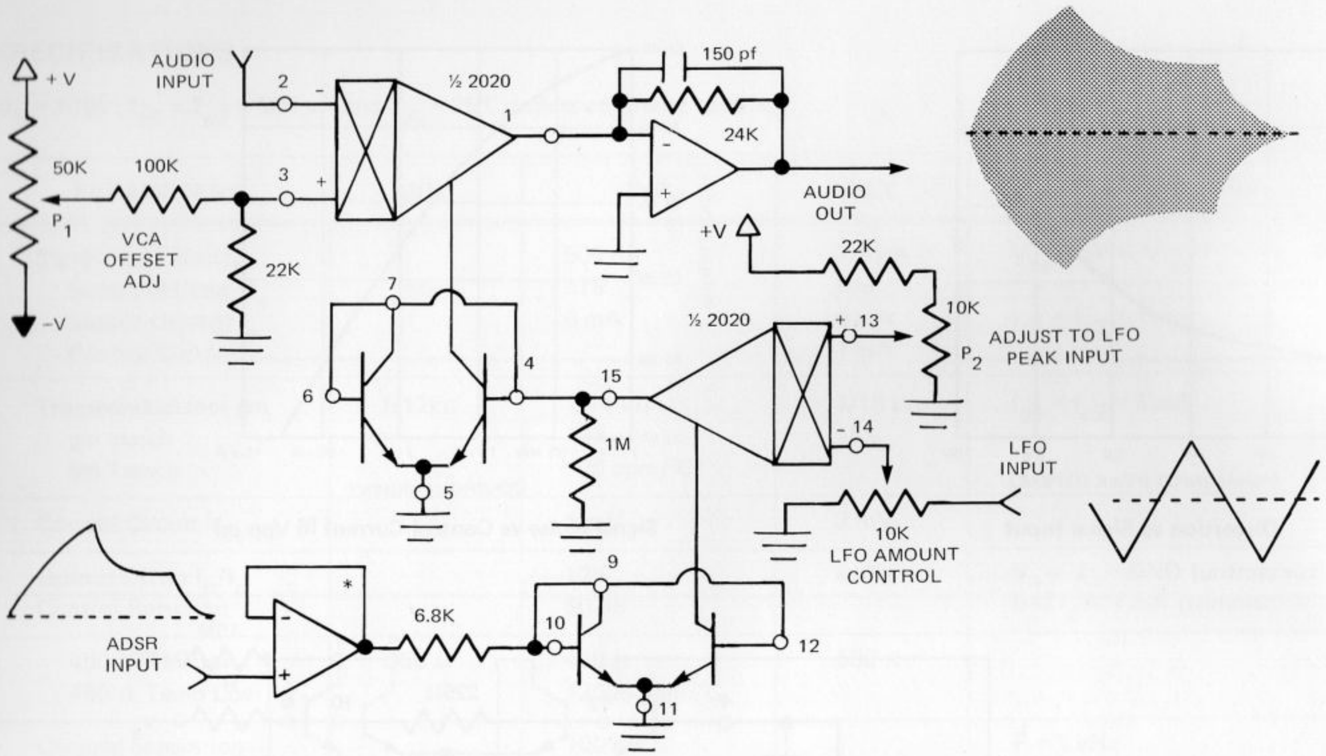
Voltage Controlled Amplifier

VCA Design Equations

$V_{out} = I_c\ V_{in}\ 15K/14$ Volts

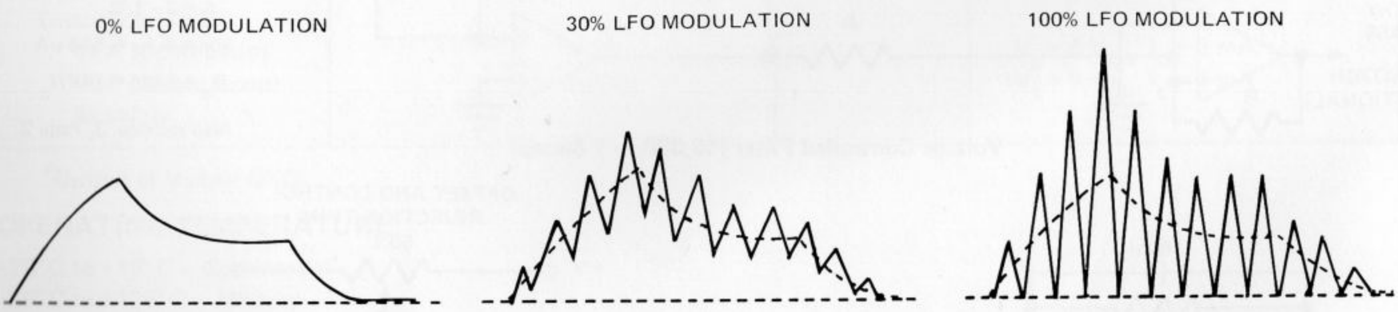
$I_c = (+V/R_{ref} + V_L/R_1)\ e^{-V_b q/KT}$

$Vb = Ve\ 450\Omega / (R_1 + 450\Omega)$



A single 2020 can be connected to add tremello to the output envelope of a VCA. Tremello depth can be controlled from 0 to 100%. The adjustment procedure is to apply an audio input to the VCA, turn the LFO amount control all the way on and gate and hold on the ADSR. Trimpot P_2 is then adjusted to give 100% modulation at the VCA output. The VCA offset trimpot P_1 is then adjusted to center the VCA output about ground.

VCA Control Envelopes



*Buffer required if driving 6.8K load from output of SSM 2050. Not needed for SSM 2055.

Solid State Micro Technology for Music

SSM2000 SSM2020 Application Note

DUAL TWO-QUADRANT MULTIPLIERS

by Ron Dow, V.P. Engineering, SSMT

LINEAR-ANTILOG VOLTAGE CONTROLLED AMPLIFIERS

The SSM 2000 and 2020 are dual two quadrant multipliers designed to be used with op amps in a wide variety of precision audio frequency applications. Each channel of the two devices has differential signal inputs, current output and its own control circuit which allows the designer an independent choice of linear or exponential control characteristics and scale factor. Both parts are fully temperature compensated and have excellent distortion and signal to noise figures.

In order to show as many applications of these versatile circuits as possible in this short space, control circuits are presented in a separate section and are referred to in the application section.

CONTROL CIRCUITS SSM 2000

To produce a linear control of amplification in the SSM2000, a linear voltage to current converter is used to supply an output control current from an input voltage. The circuit in Fig. 1A is used for a positive control voltage ($V_C/R_C = I_C$), and Fig. 1B for a negative voltage ($-V_C/R_C = I_C$). The resistor in these circuits should be chosen so that the maximum desired input voltage will produce an output current of 1mA or less. The Fig. 1A circuit has the advantage of higher input impedance and in Fig. 1B a larger control voltage can be used without running into the bias voltage on the control pin (V_{cc} −4V). The circuit in Fig. 1c will gang the gain of both halves of the I.C. to the same control voltage, without using two op amps. The 3.3K resistors force a match in the control currents in both VCA's and also offer current limiting. This approach will closely match the gain of the two channels as the voltage drop across the matching resistors is 50mV or greater.

ANTILOG CONTROL GRAPH

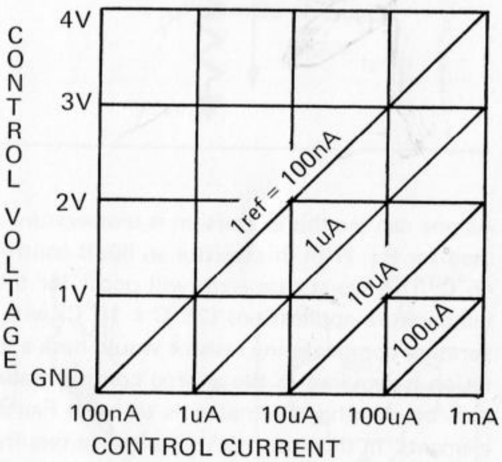


Figure 1a

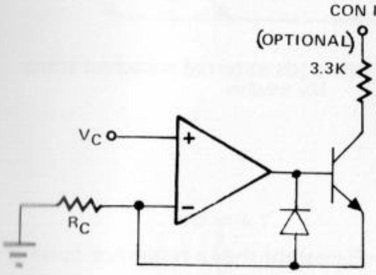


Figure 1b

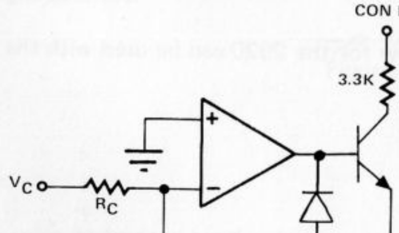
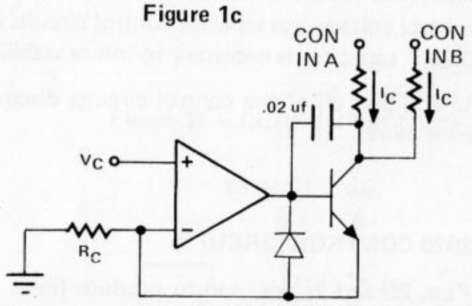


Figure 1c



Antilog control of gain is achieved by the exponential voltage to current converter in Fig. 1d. The graph shows the Antilog I_C to V_C relationship at various reference currents. As can be seen from the graph, the range of control is determined by the reference current which is set up by the circuits in Figs. 1e or 1f; 80dB for I_{REF} = 100nA etc. The scale factor, 1 decade per volt, is determined by the attenuator which in this case gives 60 mV at the op amp non-inverting input for 1 volt at the control input. In the more general case:

$$\ln(I_C/I_{REF}) = R_1 \, V_{in} / [(R_1 + R_2)\, V_T] \qquad V_T = kT/q = 26\text{mV} @ 25°\text{C}$$

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Figure 1d

Figure 1f

Figure 1g

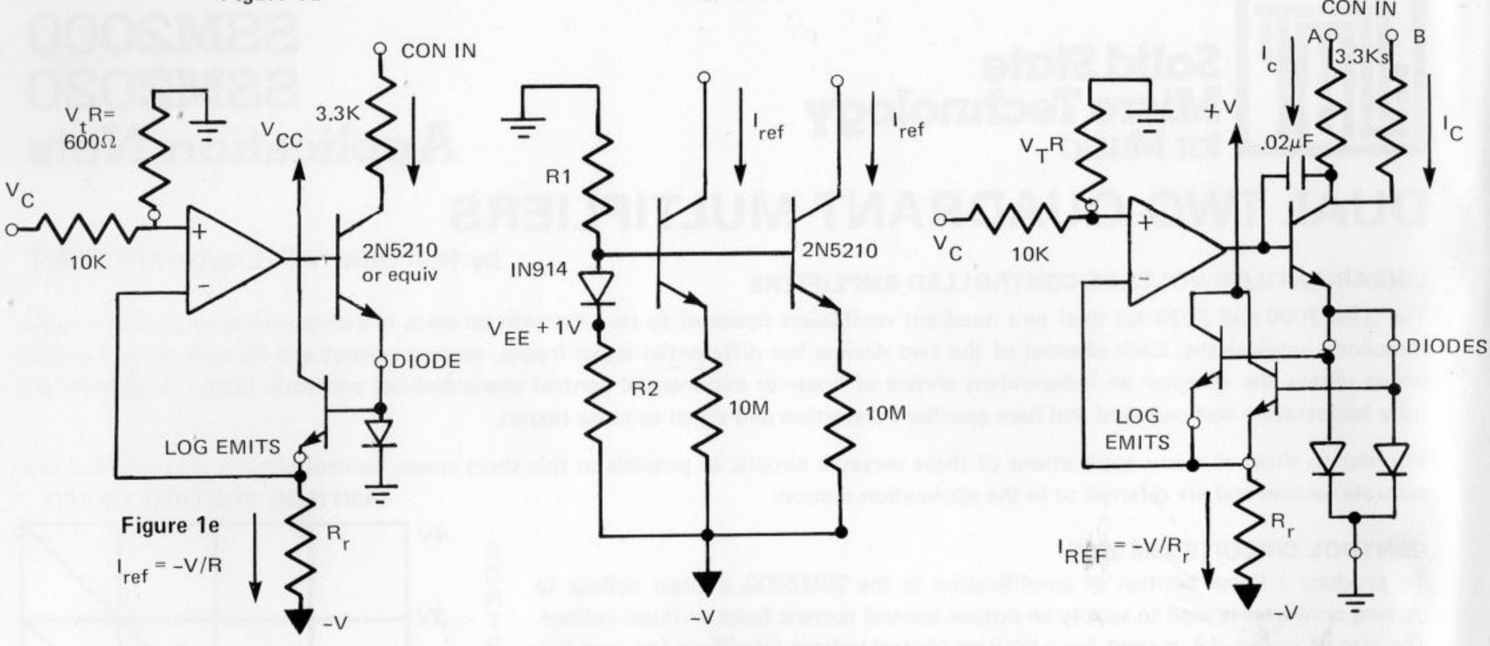


Figure 1e

As one can see this expression is temperature sensitive. This effect can be largely cancelled if the V_T resistor provided on the chip is used for R_1. With this resistor an 80dB control range can be used with a ±10% error due to temperature over the span from 10°C to 45°C. This worst case error will occur for the largest control current in the range. An ordinary resistor can be used as R_1 for room temperature applications (25°C ± 10°C) with only a 10% error if the control range is restricted to 40dB. (100 to 1). An ideal temperature compensating resistor would have a temperature coefficient of 3300ppm/°C. If extremely precise V_T temperature compensation is required in the antilog control mode, a Tel Labs Type Q81 resistor or equivalent can be used in place of the on chip V_T resistor. Another alternative is to use a Fairchild μA726 Temperature Controlled Differential pair in place of the on chip logging elements. In this case, no V_T resistor is required.

If a control range of 60dB or greater is desired, a low input bias op amp such as the LF356 or the TL082 should be used in the V to I converter with a low leakage output transistor such as a 2N5210 or a 2N930. As with the linear inverting control circuit, a 3.3K resistor is placed in a series with the collector of the output transistor for current limiting. To gang both amplifiers to the same control voltage, use separate control circuits or the circuit in Figure 1g which is designed to provide a 20dB gain change per volt; the 0.2μ F capacitor is necessary to insure stability.

In addition all of the control circuits discussed below for the 2020 can be used with the 2000 if one adds external matched transistor pairs.

2020 CONTROL CIRCUITS

Figs. 2b and 2c are used to produce linear and, or exponential voltage control of gain. Resistor R_1 establishes a reference current in the circuit which, for maximum temperature stability, should be chosen at the logarithmic center of the control range. For example if one desires a control range from 500μA to 50nA the reference current would be 5μA. A control voltage and a resistor are usually used to produce I_{in}. The 450Ω resistors shown in the figures are on the 2020 and have a temperature coefficient of around +2000ppm/°C. This largely compensates for the T factor in the exponential control equation. Due to variation of the 450Ω resistor value with processing, some selection or trimming of the input resistor may be required to produce the desired scale factor.

Linear control of gain with cutoff can be achieved by grounding both bases and eliminating resistor R_1. The control current will then be equal to V_L/R_2 with I_C going to zero with V_L at ground or below.

Fig. 2d shows both VCA control circuits ganged to the same control voltages. With this connection the control currents in the two VCA's will match and track.

Figs. 2i and 2j show two circuits for producing a control current that is proportional to the amplitude of an AC input signal. Fig. 2i is an AC to DC converter that has equal attack and decay times. Fig. 2j is a peak detector that has fast attack and slow delay. Both circuits are useful in compander and AGC applications. Inexpensive dual op amps such as the 1458 can be used without trimming for good results over a 50dB input range.

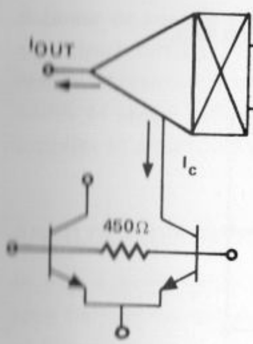


Figure 2a

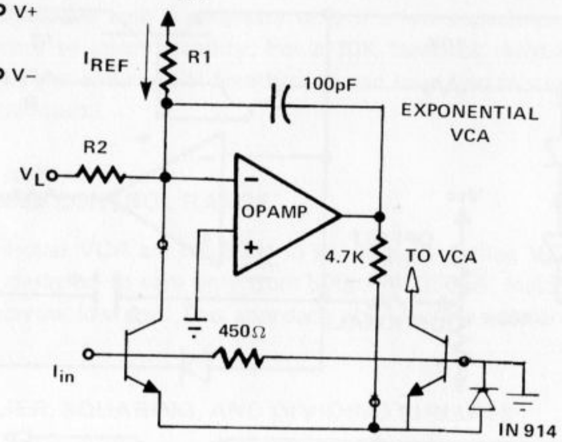


Figure 2b

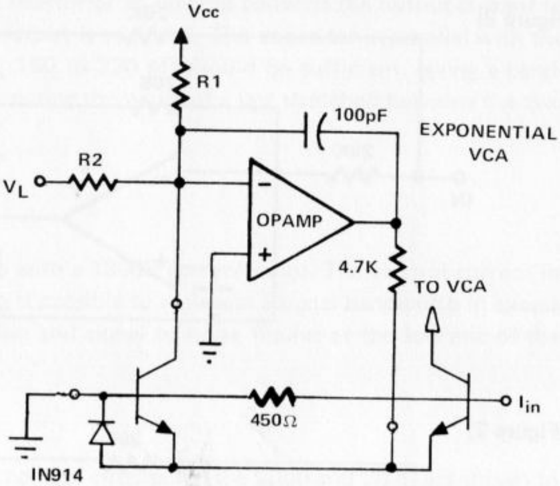


Figure 2c

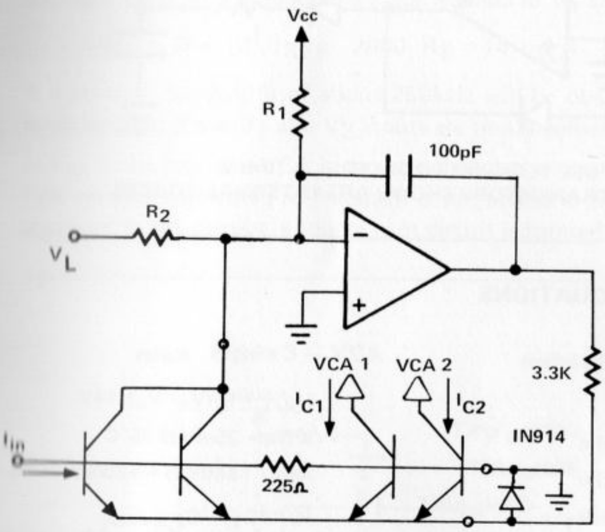


Figure 2d

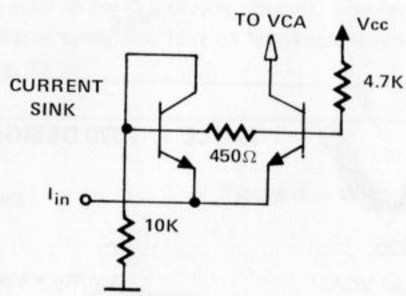


Figure 2e

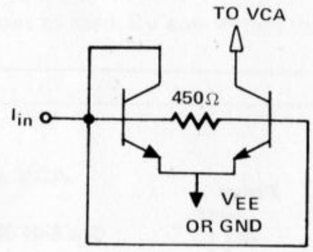


Figure 2f – CURRENT SOURCE

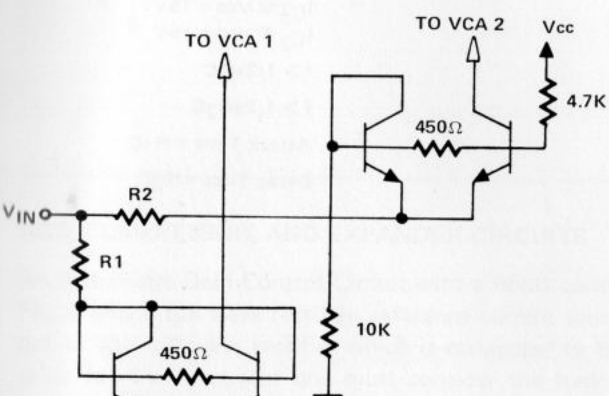


Figure 2g – VC PAN OR MIX

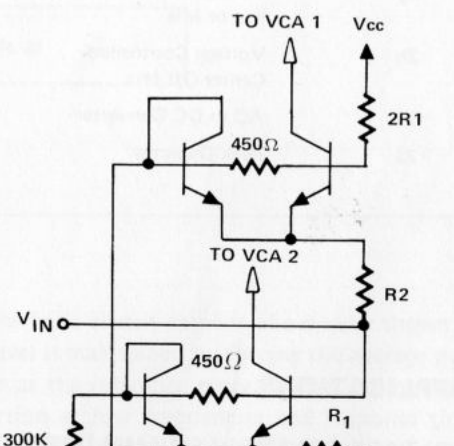


Figure 2h – VC CENTER OFF MIX

Figure 2i

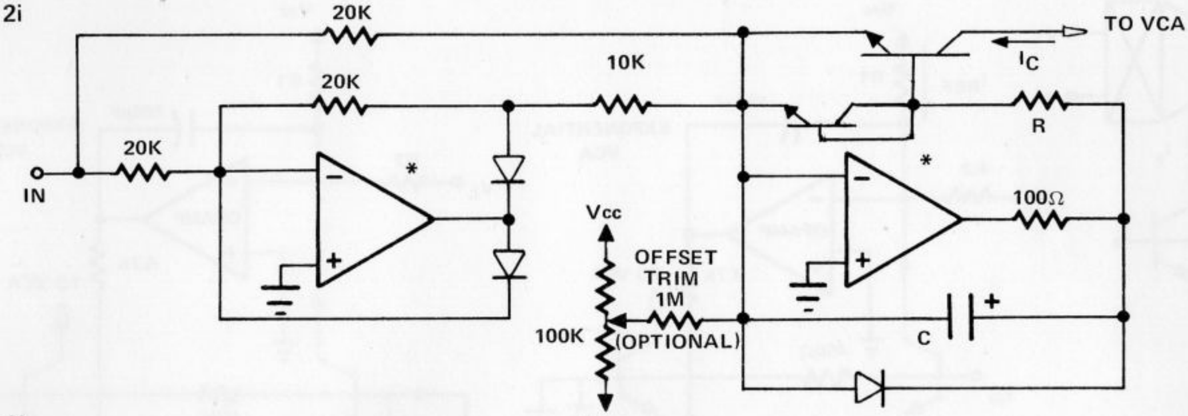
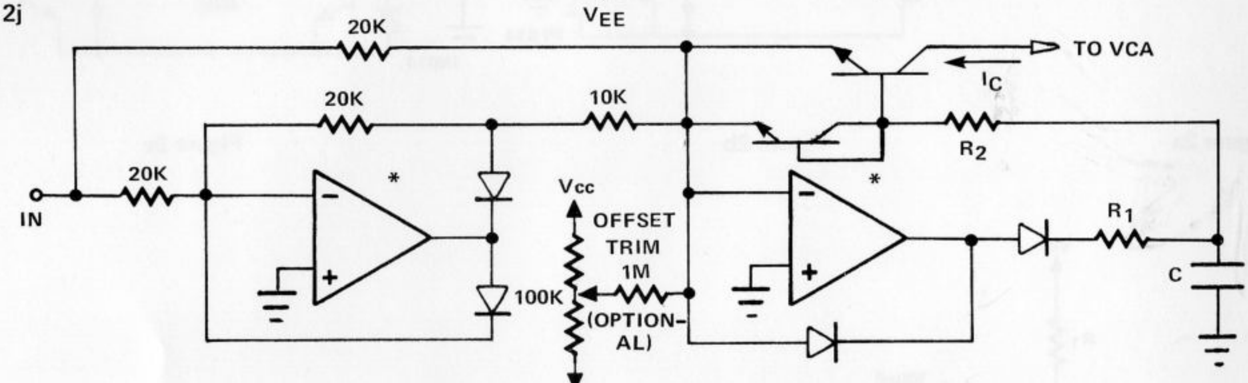


Figure 2j



*USE TL082 OR EQUIV IF $f_{in} \geq$ 10KHz
TRANSISTORS SHOWN ARE INTERNAL TO 2020.

TABLE 1. 2020 DESIGN EQUATIONS

Figure	Function	Control Current	Notes
2a	One Side SSM 2020	I_C	$I_{OUT} = I_C(V_+ - V_-)/14V$
2b	Linear-Exponential VCA	$(V_L/R_2 + Vcc/R_1)\ e^{-450 I_{in}\ q/KT}$	$KT/q = 25mV$ @ 25°C
2c	Linear-Exponential VCA	$(V_L/R_2 + Vcc/R_1)\ e^{450 I_{in} q/KT}$	450Ω TEMPCO = +2000ppm/°C
2d	Ganged Linear-Exponential VCA	$1/2\ (V_L/R_2 + Vcc/R_1)\ e^{-225 I_{in} q/KT}$	$I_C = I_{C_1} = I_{C_2}$
2e	Current Sink	$-I_{in}$	
2f	Current Source	I_{in}	
2g	Voltage Controlled Pan or Mix	$(V_{in} - 0.7)/R_1$ $(9.3 - V_{in})/R_2$	I_{C_1} @ Vcc = 15V I_{C_2} @ Vcc = 15V
2h	Voltage Controlled Center-Off Mix	$3(V_{in} - 5.7)/2R_1$ $(4.3 - V_{in})/R_2$	I_{C_1} @ Vcc = 15V I_{C_2} @ Vcc = 15V
2I	AC to DC Converter	Vpp in/40K	$f > 1/2\pi RC$
2J	Peak Detector	Vpp in/40K	$f > 1/2\pi R_2C$ Attack Time = R_1C Decay Time = R_2C

APPLICATIONS

VOLTAGE CONTROLLED AMPLIFIER

Fig. 3 shows half of a dual VCA used in its principle application. The output voltage is related to the input voltages and control current by:

$$V_{OUT} = I_C\ (V_s{+} - V_s{-})\ R_F/V_E \qquad V_E = 11.8V \text{ for 2000, } V_E = 14V \text{ for 2020}$$

A linear or antilog voltage to current converter sets the control current and a resistor or an op amp converts the output current to an output voltage. The output op amp is necessary only if a low-impedance output is required. The capacitor in parallel with the feedback resistor is necessary to insure stability. For a 10K feedback resistor, 150 to 220 pF should be sufficient, giving a bandwidth of about 100kHz. Offset and control feedthrough can be nulled by connecting the wiper of a pot stretched between the two supplies to an unused signal input.

ANTILOG VCA WITH 130dB CONTROL RANGE

In Fig. 4, both halves of a dual VCA are cascaded to produce an antilog VCA with a 130dB control range. The control current in each half of the device is designed to vary only from 500μA to 250nA, making it possible to maintain a signal bandwidth in excess of 50kHz down to the extreme low end. This approach also improves distortion and signal to noise figures at the low end of the control range.

4 QUADRANT MULTIPLIER, SQUARING, AND DIVIDING CIRCUITS

In Fig. 5 half of a VCA is used as a four quadrant multiplier. The appropriate control circuits for the 2000 and 2020 are shown in the figure. The output current into a virtual ground is:

$$I_{OUT} = -V_1 V_2/(R_C V_{ee})$$

To adjust the circuit for proper operation, a signal is applied to the V_2 input with V_1 grounded and R_p trimmed for minimum feedthrough. V_2 is then grounded, a signal applied to V_1 and R_F trimmed for minimum feedthrough which should occur when:

$$R_F = 1.18 \times 10^1/I_{REF} \quad 2000, \quad R_F = 14/I_{REF} \quad 2020$$

A maximum bandwidth of about 250kHz will be obtained with $I_{REF} = 200\mu$A. The op amp converts the output current to a buffered voltage. If the V_1 and V_2 inputs are tied together the output will be the square of the common input.

In Fig. 6 one half of a VCA is shown connected with an op amp to form a divider circuit. The output is proportional to V_1/V_2. The 22M resistor connected in feedback is optional and prevents a complete loss of feedback when I_C goes to zero. By connecting the V_2 input to the output, a square root circuit is formed. (Fig. 7)

Figure 3 – VCA

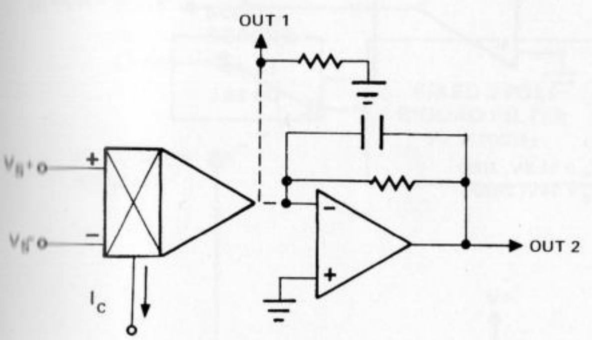
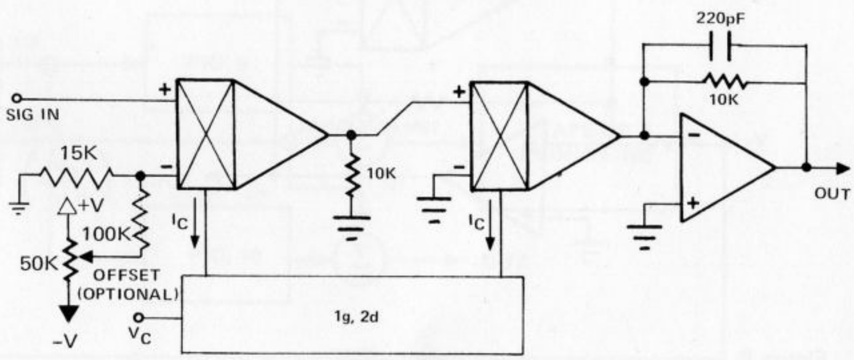


Figure 4 – Wide Range VCA



AGC, COMPRESSOR, AND EXPANDER CIRCUITS

An Automatic Gain Control Circuit with a 60dB control range is shown in Fig. 8. The basic circuit consists of a divider shown in Fig. 6 plus a full wave rectifier, reference current source and integrator. The output level is maintained by forcing the average output of the full wave rectifier which is connected to the signal output to be the same as the reference current. In choosing the C value for the integrator one must consider the tradeoff between modulation distortion at low frequencies and response time to change of input level. The circuit shown is optimized for the 2020 but the 2000 can also be used by reversing the direction of the diodes and reference current source in the full wave rectifier and by connecting the 22K integrator output resistor to the Con In pin of the 2000. An AGC with up to a 100dB control range can be made by cascading two divider circuits and ganging the two VCA sections to the output of the integrator. (Extreme care must be taken in layout to bypass supplies and shield the summing modes of the dividers from stray signals.)

Figure 5 – 4 Quadrant Multiplier

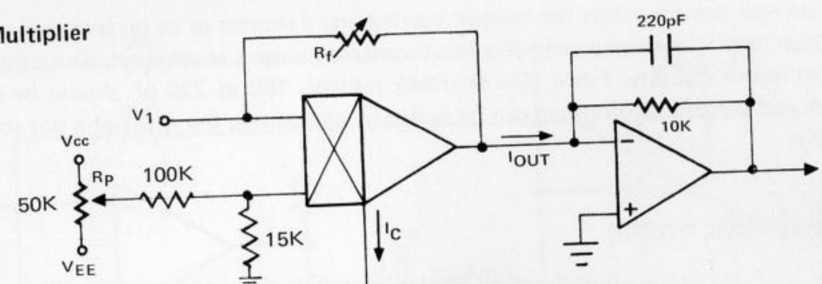
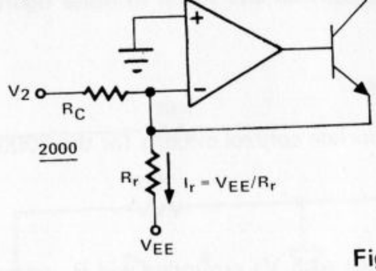


Figure 5a



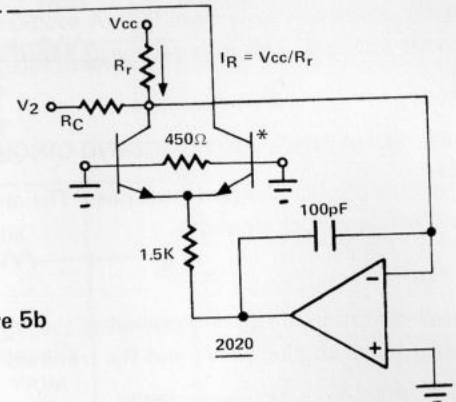
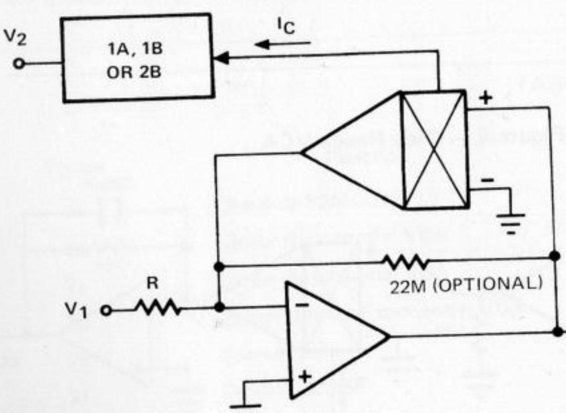


Figure 5b

Figure 6 – Divider



$$V_{OUT} = \left(\frac{V_e\, V_1}{R\, I_C}\right) \quad \begin{array}{l} V_e = 11.8V,\ 2000 \\ V_e = 14V,\ 2020 \end{array}$$

Figure 7 – Square Root Circuit

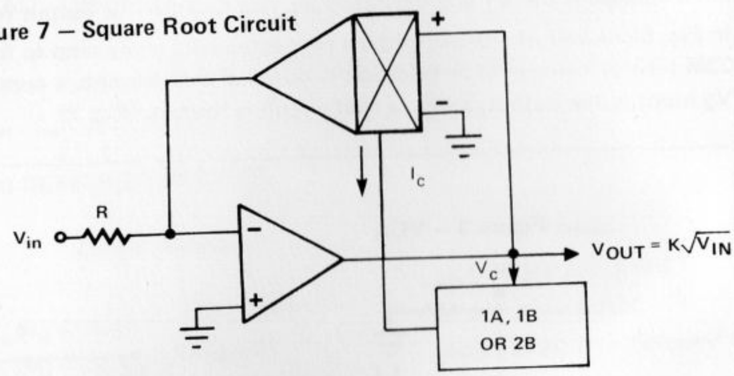
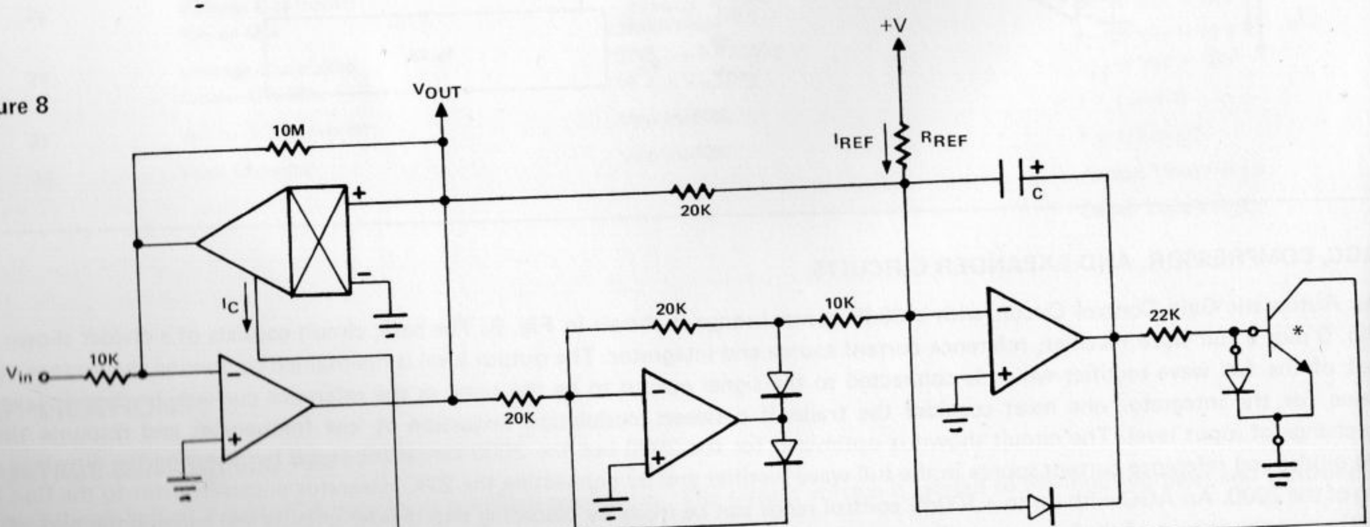


Figure 8



$V_{OUT}\,PP = I_{REF}\ 20K$ FOR $f > 1/\,[2\pi C\ (6.7K\Omega)]$

*INTERNAL TO 2020 SEE FIGURE 2F

Figure 9 — Compressor

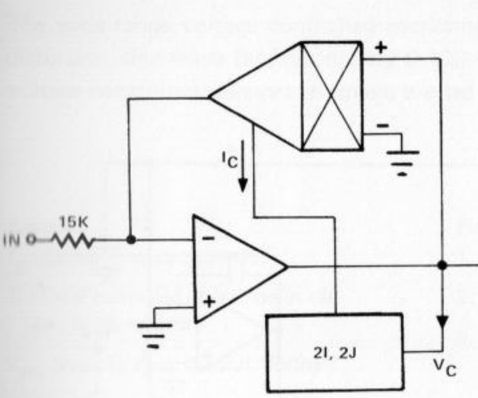
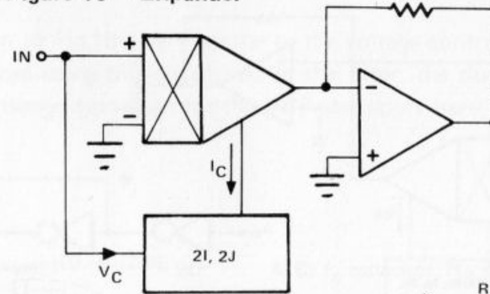


Figure 10 — Expander



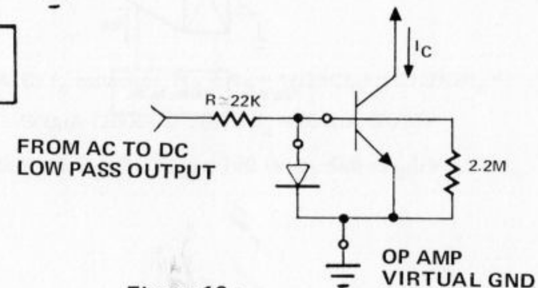


Figure 10a

Figs. 9 and 10 show a VCA being used in a compressor and expander circuit respectively. A compressor is a circuit that changes its output by 6dB in response to a 12dB input signal level change. An expander's output level changes by 12dB in response to a 6dB level change at the input. These functions are reciprocal and can be used in noise suppression systems when sending signals to and from a noisy media like a phone line or magnetic recording tape. The block diagram of such a system is shown below in Fig. 11. Since the most objectionable audio noise occurs in the band from 500Hz to 5kHz, noise suppression need only be applied to this band of frequencies. The signal to or from the media is divided by a filter into lows, which are unaffected and highs which are companded. This enables one to choose filter capacitors in the AC to DC control circuits for the compressor and expander which are small enough to give good dynamic performance in response to a sudden level change while maintaining an adequate distortion figure for all audio signal components. Figure 10a shows a 2.2M resistor which can be added to the expander control circuit to produce a noise gate for the highs when the signal level falls below -60dB.

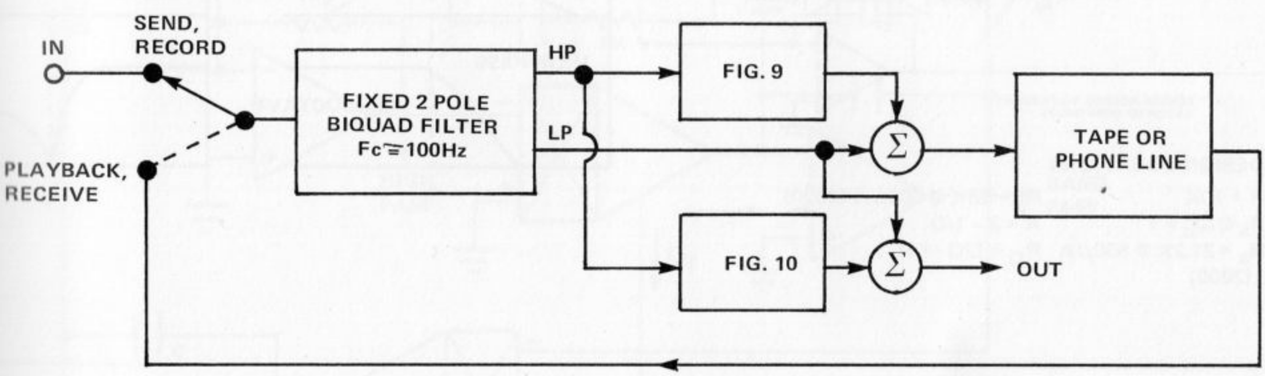


Figure 11

MIXERS, FADERS AND PANNING CIRCUITS

Fig. 12 is the basic circuit for voltage controlled mixing. This circuit can be expanded to accomodate more inputs by connecting additional VCA outputs to the summing mode of the op amp. Two Fig. 2b control circuits can be used to independently control the mix level from the two inputs. If control is ganged by using circuits 2g or 2h, a voltage controlled pan or center-off mix can be implemented.

Stereo panning can be accomplished by using two Fig 12 circuits and controlling one with the inverted control signals of the other. This will produce an effect where the right and left sound sources will change sides with an equal monoral mix in the center. With two more Fig. 12 circuits this concept can be extended to quadraphonic systems for interesting spacial and rotating sound effects.

Fig 13 is a control circuit for an automated mixer-fader using Fig 12. With the digital input high, only the A input appears at the output under exponential control of V_1. When the logic input goes low the A level decays to zero and the B level attacks to a level controlled by V_2 with a time constant ½ RC. The transition between the two sound sources is smooth and pleasant to the ear.

Figure 12

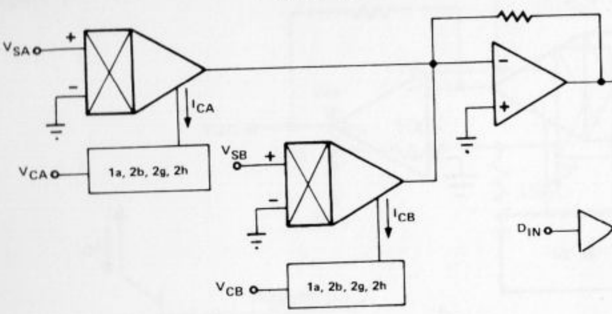
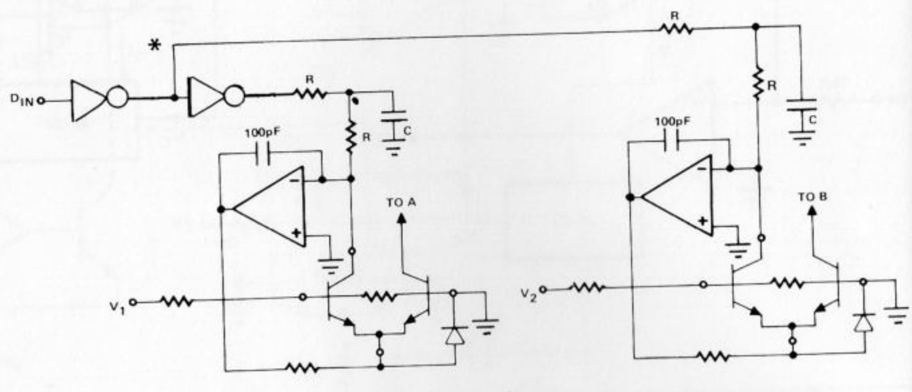
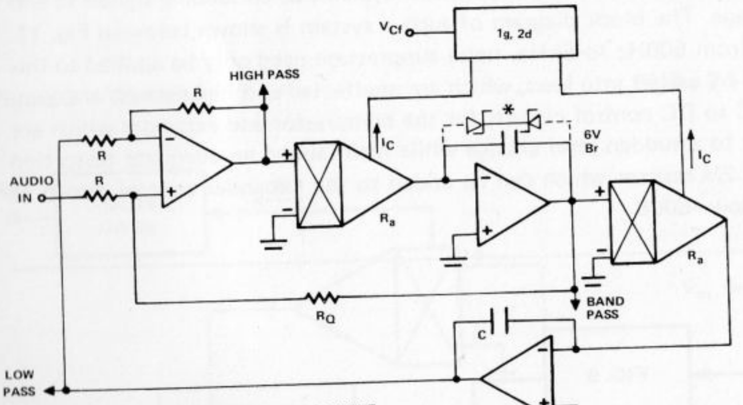


Figure 13



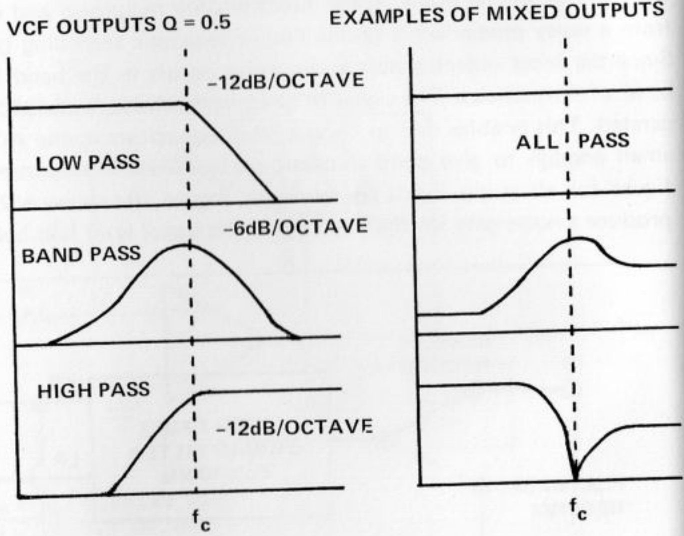
* ½ CMOS 4007 WITH V_{CC} = +15V

Figure 14



• ZENERS NEEDED TO PREVENT LATCH UP (2000 ONLY)

Table 2. VCF Outputs & Mixed Outputs



DESIGN EQUATIONS

$R = 10K$

$R_a C \omega_o = 1$

$R_a = 21.2K$ @ $500\mu A$ (2000)

$R_a = 28K$ @ $500\mu A$ (2020)

$A = 2 - 1/Q$

$R_Q = (2Q - 1)R$

VOLTAGE CONTROLLED FILTERS AND EQUALIZERS

Using the circuit in Fig. 14 a voltage controlled filter with a 10,000 to 1 control range or more can be implemented. The low and high pass outputs have 12dB/octave rolloffs and the band pass output has 6dB/octave skirts. Such circuits can be seried to produce more complex filters. In this application a dual VCA can be thought of as a pair of matched voltage controlled resistors. The R_a in the design equation for the cutoff frequency given next to the figure is 21.2K for the 2000 and 28K for the 2020 when the control current in both halves of the device is 500 μA. At 50 μA control current, the value of the R_a will increase by a factor of 10 and so forth. For control ranges of 1000 to 1 or greater, low input bias op amps should be used in the control circuit and signal section of the filter. The LF356 and the TL084 quad op amp perform well offering low input bias, low noise and wide power bandwidth. If antilog control is used, the greatest control accuracy for a 10,000 to 1 range is obtained for control currents of 500 μA to 50 nA. VCA's used in an output mixer with fixed or voltage controlled filters can implement a remote-controlled or automated equalizer. Fig 15 is a VCF that has, in addition to cutoff frequency control, a Q that can be voltage controlled from less than 0.5 to 250. This filter has the additional feature that the gain remains constant at max pass. Low input bias wide band-width op-amps must be used in this circuit to achieve the stated performance over the entire audio frequency range.

VOLTAGE CONTROLLED QUADRATURE OSCILLATOR

The wide range voltage controlled oscillator circuit shown in Fig.16 is very similar to the voltage controlled filter. An extremely low distortion sine wave (approximately 0.1%) can be produced using this circuit. As in the filter, the dual VCA is used as a pair of voltage controlled resistors to tune a biquad stage. All the design tips given for the VCF also apply here. The design procedure is:

Specify:

f_o (Center frequency in Hz)

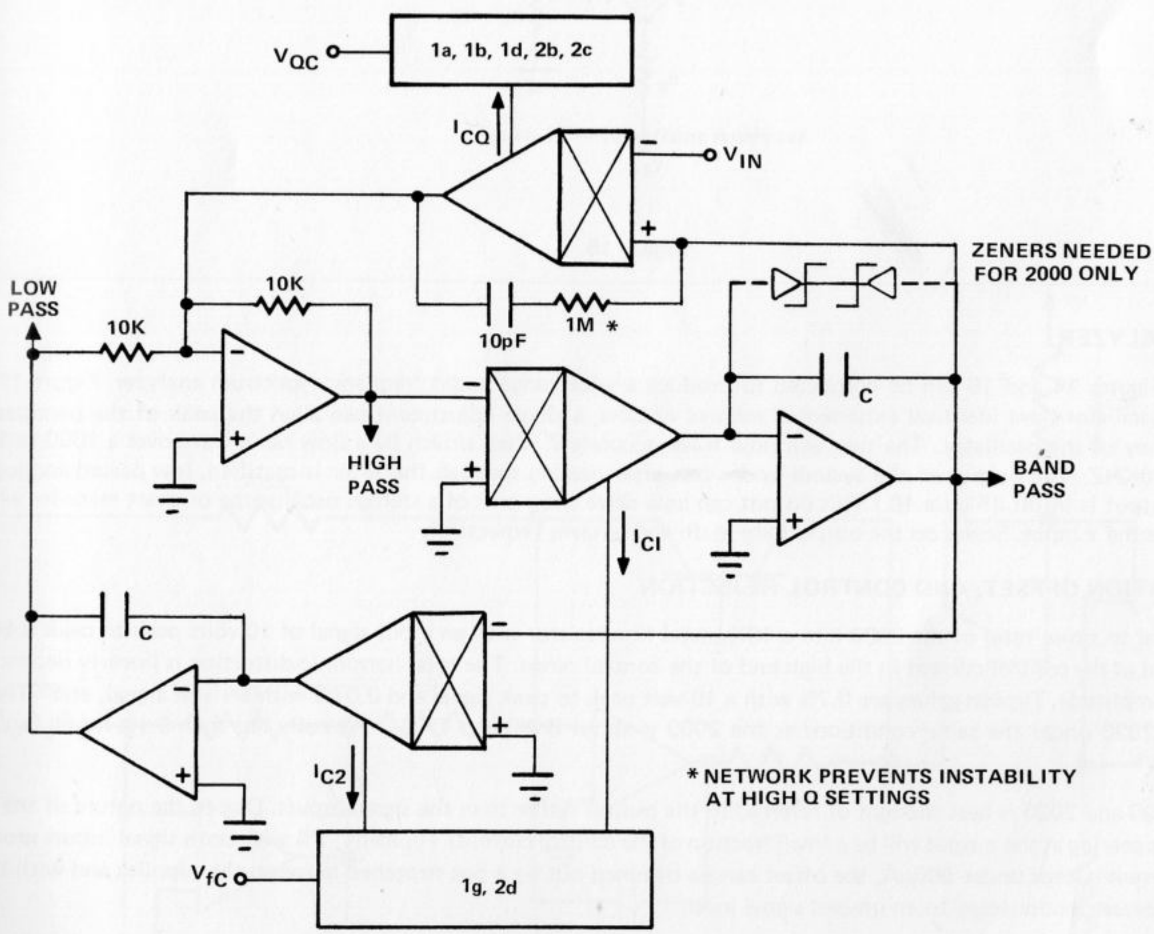
D (Total harmonic distortion in dB. i.e., 0.1% = 60dB

V_{po} (Peak to Peak Output Voltage)

Find:

1. Transfer ratio; $r = 10^{((D-9)/20)} = 8Q$
2. Peak limiter voltage; $V_{lp} = 2V_{po}\pi/r$
3. Q setting resistor; $R = r/16\pi Cf_o$
4. C: f_o capacitor; $R_2 = R_3 = 1/(2\pi Cf_o) = 21.2K@I_c = 500\mu A$ (2000) or $28K @ I_c = 500\mu A$ (2020)
5. R_7 & R_8; $R_7 = R_8 = 100\ (V_{CC} - 0.6 - V_{lp})/V_{lp}$

FIGURE 15



For other filters that can be used with the 2000 and 2020 see "Active Filter Cookbook", Don Lancaster; Howard W. Sams and Co.

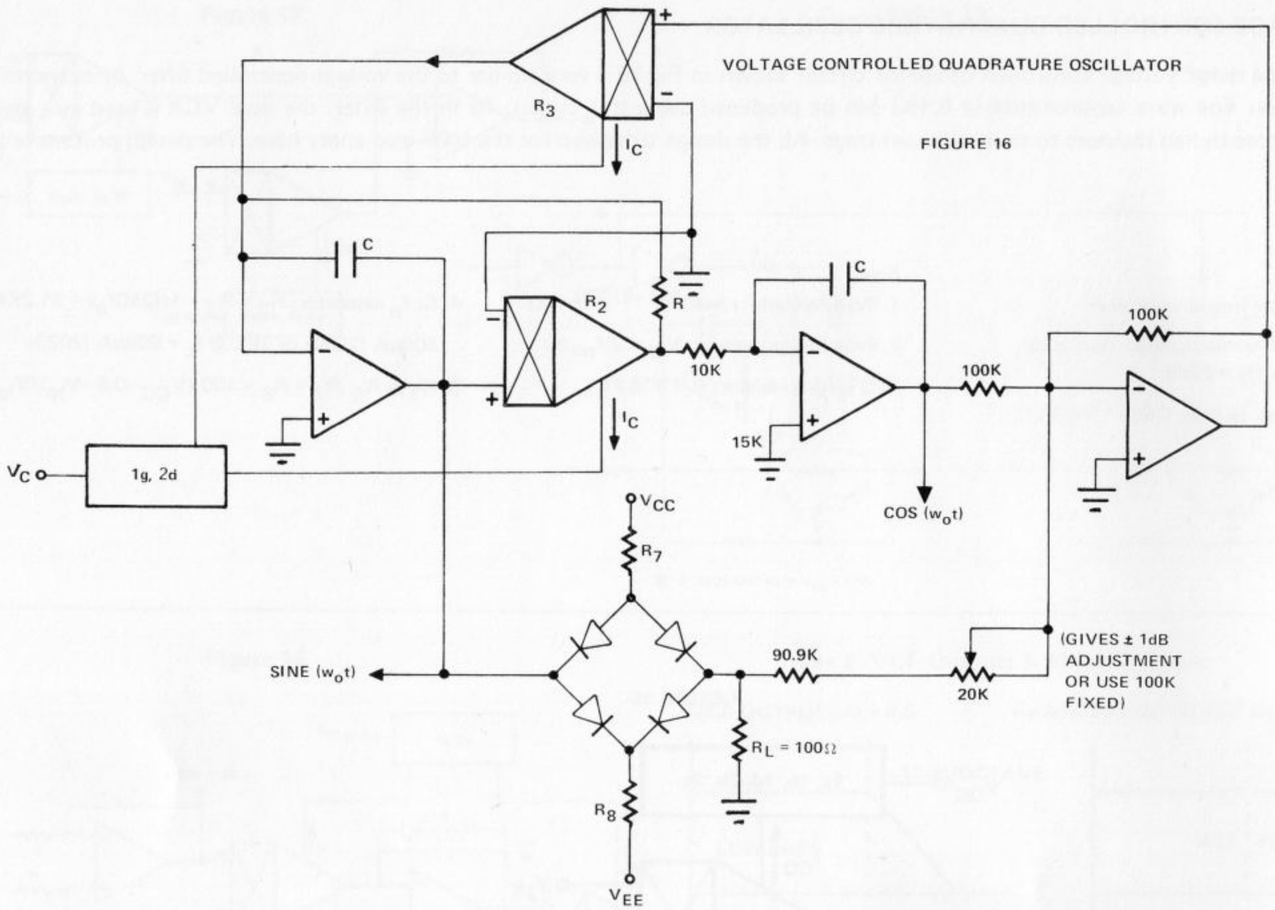


Figure 16

SPECTRUM ANALYZER

The circuits in Figures 14 and 16 can be combined to produce a wide range audio frequency spectrum analyzer. Figure 17. Since both filter and oscillator have identical exponential control circuits, a single adjustment can align the peak of the bandpass filter with the frequency of the oscillator. The two will now track accurately when driven by a slow ramp wave over a 1000-to-1 sweep from 20HZ to 20KHZ. The output of the system under test after passing through the filter is rectified, low passed and logged so that the final output is in db. (Figure 18.) This output can now drive the y axis of a storage oscilliscope or chart recorder while the ramp wave drives the x input. Scales on the outputs are in db and octaves respectively.

NOISE, DISTORTION,OFFSET, AND CONTROL REJECTION

The output signal to noise ratio of the 2000 into a 10K metal film resistor with an input signal of 10 volts peak to peak is 80dB almost independent of the control current in the high end of the control range. The total harmonic distortion is linearly dependent on the input signal amplitude. Typical values are 0.7% with a 10 volt peak to peak signal and 0.07% with a 1 volt signal, etc.* The signal to noise of the 2020 under the same conditions as the 2000 is about 86dB and THD is typically less than 0.1% for all input levels less than 6 Vpp.

Offset in the 2000 and 2020 is best thought of referred to the output rather than the signal inputs. Due to the nature of the design, the D.C. offset appearing at the output will be a small fraction of the control current: typically, 2% with both signal inputs grounded . If the control current is kept under 500μA, the offset can be trimmed out by a pot stretched between the supplies and with the wiper feeding an attenuator connected to an unused signal input.

Offset and control rejection are related and both can be trimmed out with the same adjustment. Control rejection for the 2000 is typically 20dB untrimmed and 46dB trimmed. The corresponding figures for the 2020 are 24dB and 56dB respectively.

*Note: If the circuits in figures 1C or 1G are used as control circuits to gang both sides of the SSM2000 to the same control voltage, the distortion figure is doubled and includes some second harmonic distortion. Using two control circuits with a common input will avoid this.

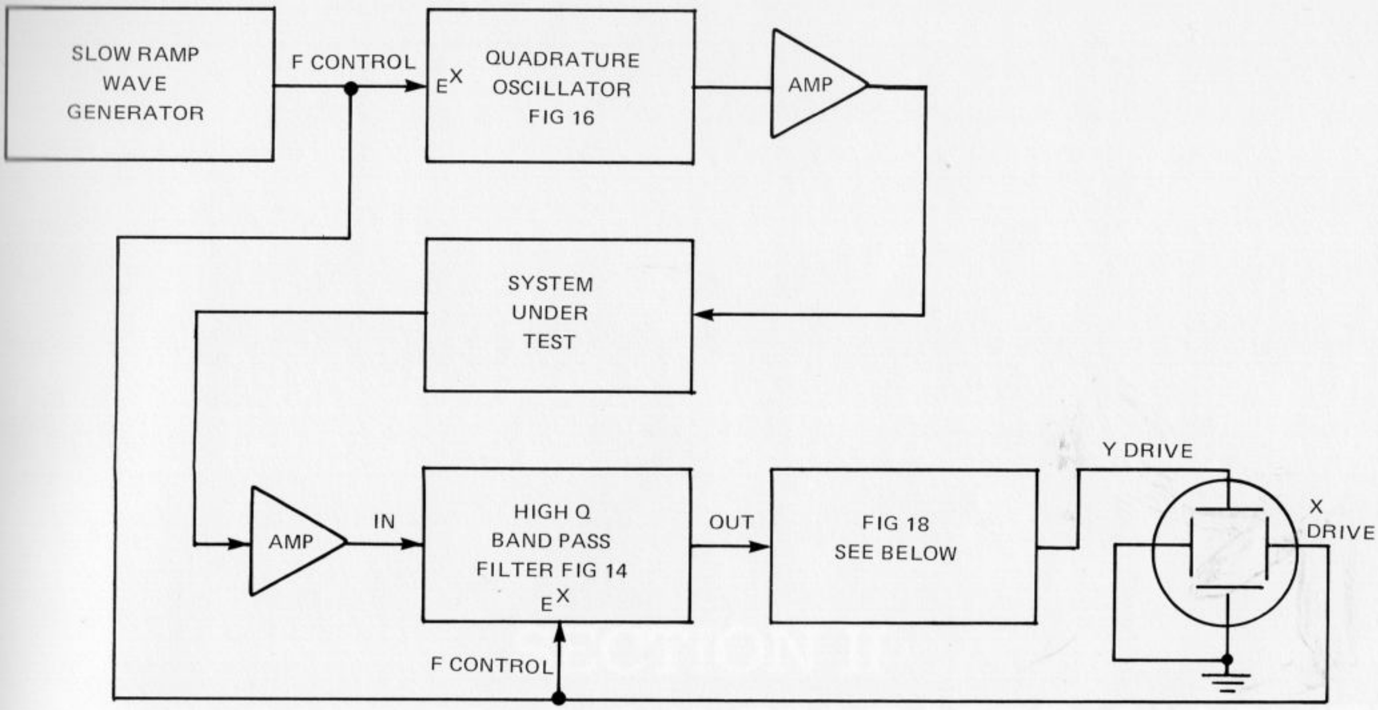


Figure 17 — Spectrum Analyzer

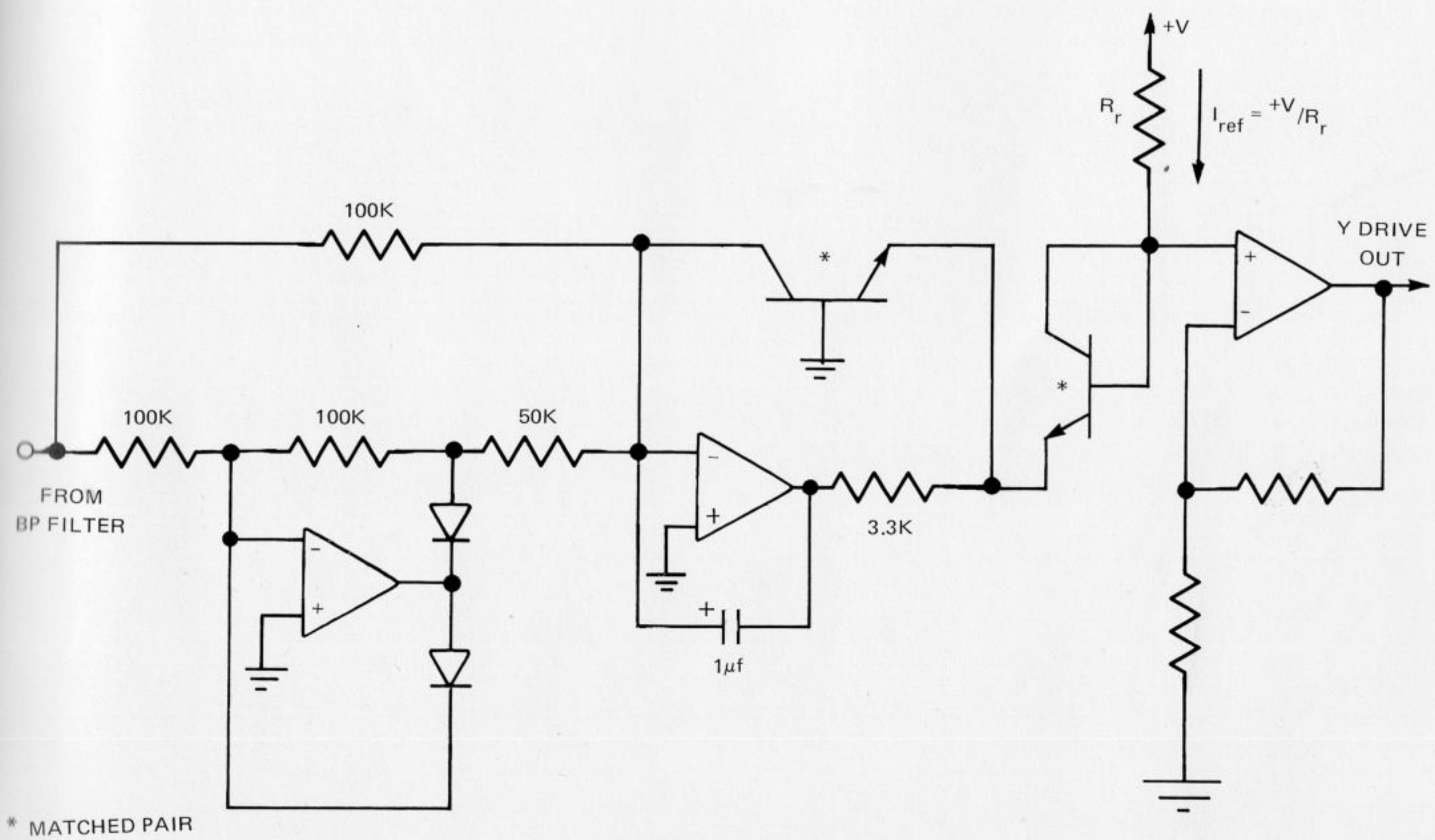


Figure 18 — Output Conversion Circuit

SECTION II
ELECTRONIC MUSIC INTEGRATED CIRCUITS

Solid State Micro Technology for Music

SSM 2030

VOLTAGE CONTROLLED OSCILLATOR

SSM2030 DESCRIPTION

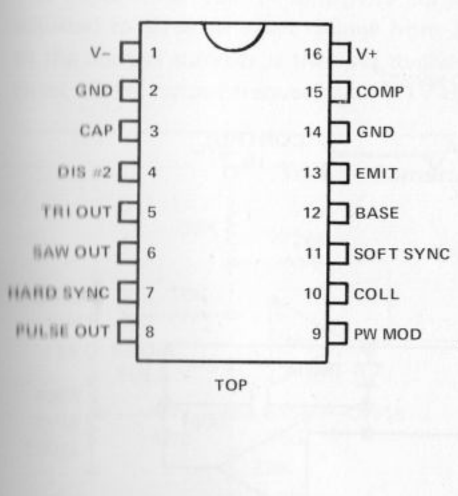
The SSM2030 is a precision voltage controlled oscillator designed specifically to meet the waveform and accuracy requirements of electronic music systems. It has both exponential and proportional linear sweep inputs which can control frequency over a 1,000,000 to 1 range with the same capacitor. Sweep accuracy is better than 0.25% over a 1,000 to 1 range and 0.1% over 100 to 1. The device has simultaneous sawtooth, triangle and pulse outputs. An internal comparator provides control of pulse output duty cycle from 0 to 100%. Hard and soft sync inputs make possible a rich variety of modulation and harmonic locking effects.

FEATURES

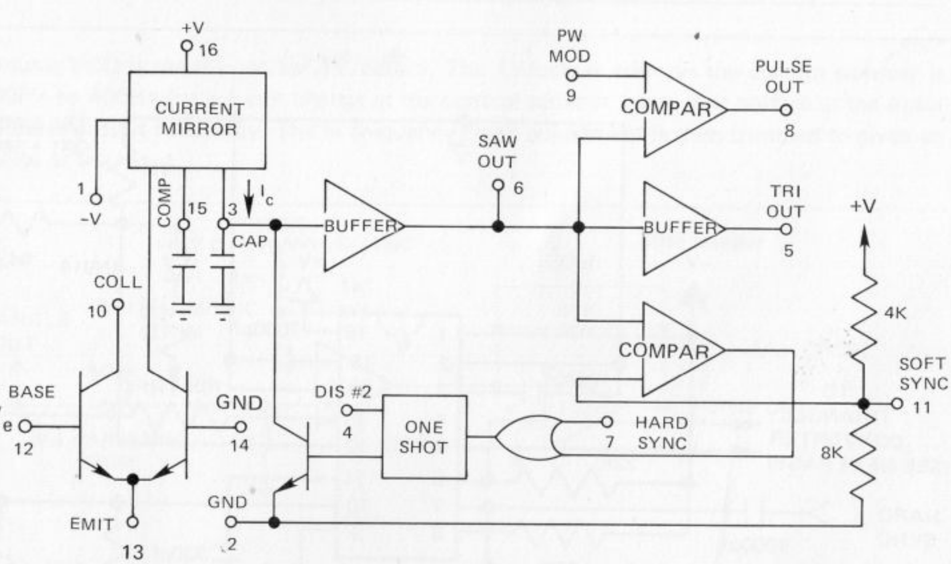
- Simultaneous Exponential and Proportional Linear Sweep Inputs
- High Sweep Accuracy (0.25% 1000 to 1)
- 1,000,000 to 1 Sweep Range
- 200 kHz Max Operating Frequency
- Simultaneous Sawtooth, Triangle and Pulse Outputs
- Pulse Duty Cycle Voltage Control Range (0 to 100%)
- All Outputs Short Circuit Protected
- Hard and Soft Sync Inputs
- Max Supplies ±18V

APPLICATIONS

- Music Synthesizers
- Electronic Organs
- Electronics Games
- Waveform Generation
- V to F and F to V Conversion
- Modulation Control Circuits
- Wide Range Phase-Locked Loops
- Frequency Multiplication and Division



Pin Diagram



Block Diagram

Solid State Micro Technology for Music, 2076B Walsh Avenue, Santa Clara, CA 95050, USA
(408) 248-0917 Telex 171189

SPECIFICATIONS

$V_s = \pm15V$ AND $T_A = 25°C$

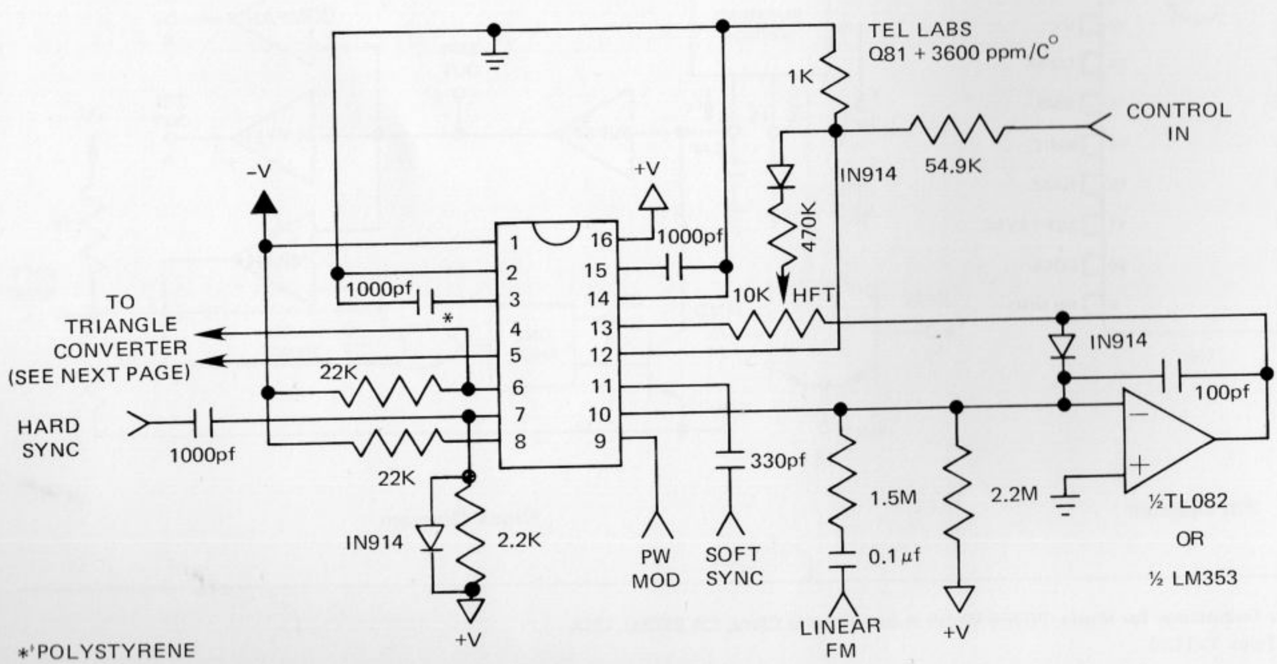
STORAGE TEMPERATURE

-55°C to +125°C

OPERATING TEMPERATURE

-25°C to +75°C

PARAMETER	CONDITIONS	MIN	TYP	MAX
V_s		±9V	±15V	±18V
Supply Current	$I_C = 1$ mA	8 mA	12 mA	16 mA
Buffer Leakage	$I_C = 0$		100 pA	1nA
Sweep Range	C = 1000pF	10^6:1	10^7:1	–
Operating Frequency	C = 1000pF	0.02 Hz	–	200kHz
Sawtooth Amplitude		9.5 Vpp	10 Vpp	10.5 Vpp
Pulse Amplitude		7.0 Vpp	7.5 Vpp	8.0 Vpp
Sawtooth Fall Time				
Buffer Output		–	500 nsec	–
Buffer Input		–	200 nsec	–
Pulse Output				
Fall Time		–	200 nsec	–
Rise Time		–	200 nsec	–
Exponential Confirmity (Trimmed)				
1000:1	20Hz-20kHz, C = 1000pF	–	0.25%	–
100:1	100Hz-10kHz, C = 1000pF	–	0.1%	–
1000:1 Oscillator Matching	20Hz-20kHz, C = 1000pF	–	0.1%	–
Linearity (Trimmed) 1000:1	20Hz-20kHz, C = 1000pF, V_e = GND	–	0.05%	–
Output Current (before clipping)				
Sawtooth Output		1.8 mA	2.4 mA	3.4 mA
Triangle Output		1.8 mA	2.4 mA	3.4 mA
Pulse Output		3.5 mA	4.6 mA	6.5 mA
Control Circuit V_{OS}	$I_e = 100\ \mu A$	–	1 mV	3 mV
Power Supply Sensitivity		–	0.5%/V	1%/V
Pulse Mod Input Bias		–	1 μA	2.5 μA
Temperature Stability	$V_{PIN\ 12}$ = GND	–	50ppm/°C	



Basic Connection

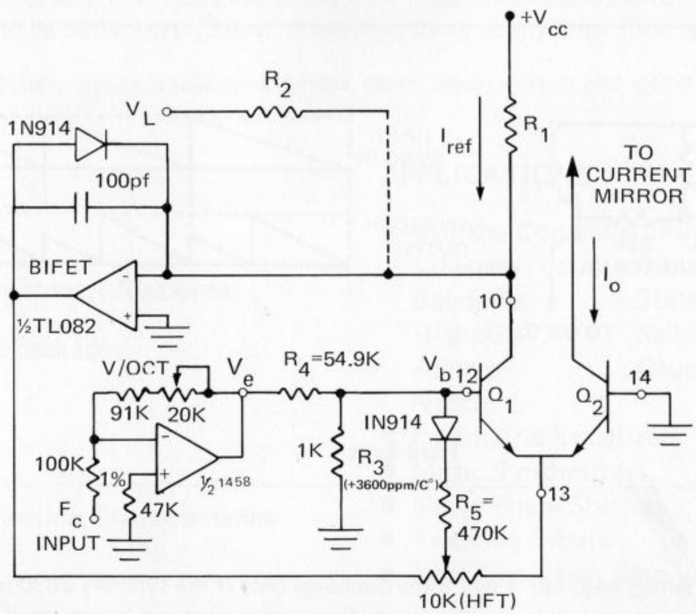
The frequency control circuit shown in the diagram is similar to many modular designs now in use. A low input bias op amp is used to force the current in Q_1, to be equal to the reference current established by R_1 and the linear FM voltage (if any). The current in the output transistor Q_2 is:

$$I_o = (V_+/R_1 + V_L/R_2)\, e^{-V_b q/KT}$$

As one can see, the term in the exponent is temperature dependent. This problem can be addressed by making V_b temperature dependent.

$$\frac{d\,KT}{dT\ \ q} = 3300\text{ppm}/^\circ\text{C @ } 25^\circ\text{C} \qquad V_b = \frac{V_e R_3}{R_3 + R_4} \qquad R_3 = 1K \quad R_4 = 54.9K$$

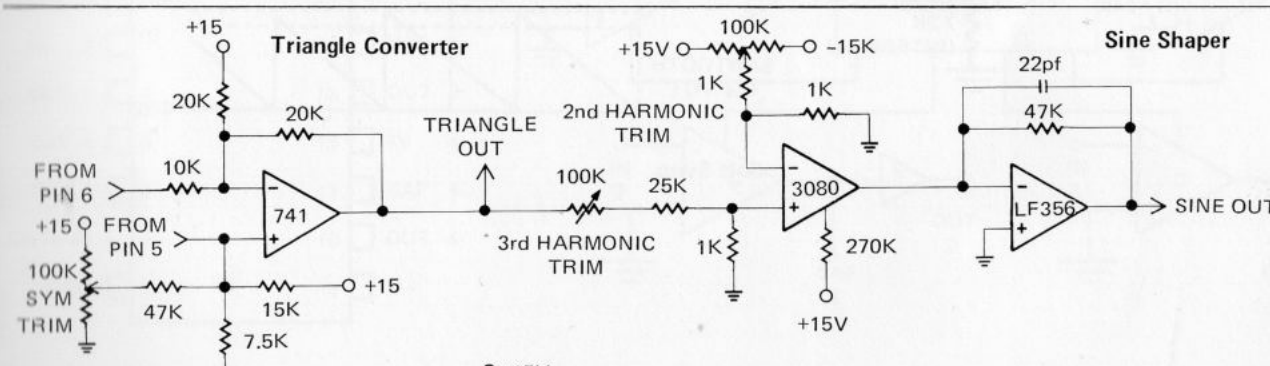
Since R_4 is large compared to R_3 one can use a Tel Lab type Q81 resistor (T.C. approximately 3600 ppm/°C) to give V_b the necessary temperature dependence. Best results are obtained with the Q81 thermally coupled to the package.

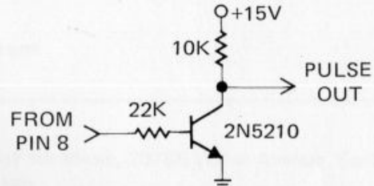


Control Circuit

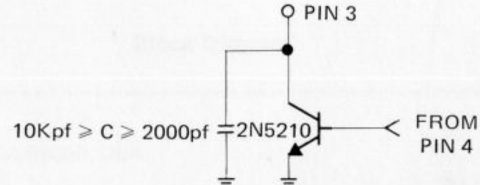
VCO ADJUSTMENT PROCEDURE

The exponential control sensitivity for a music VCO is usually set for 1V/octave. The 1V/octave trim on the control summer is adjusted to given an exact change from 200Hz to 400Hz for a 1 volt change at the control summer input. The voltage at the input to the control summer is then set to give a 5kHz output frequency. The hi frequency track adjustment is then trimmed to given an exact 10kHz output frequency for a 1V change at the input.





Large Amplitude Pulse Output Circuit



Circuit For Discharging Large Cap

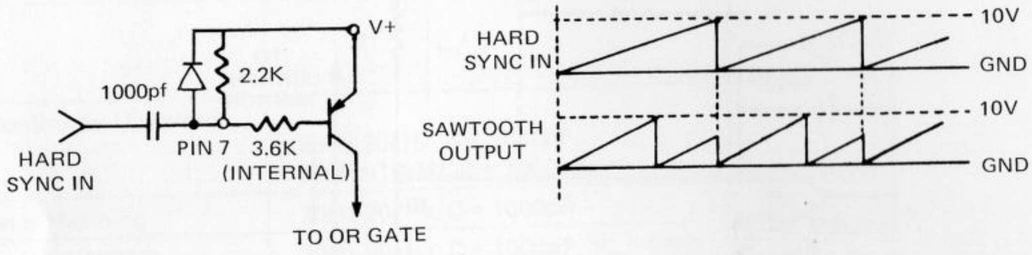
OUTBOARD CIRCUITS

The sawtooth and "triangle" output pins are both emitter followers from the internal sawtooth. The triangle waveshaping circuit (Figure 3) uses the sawtooth output and an emitter follower from the same point which is biased to give half a sawtooth. The subtraction performed by the 1458 of the sawtooth from this waveform results in a triangle output. A true 741 type op amp must be used here, as its lower slew rate ignores the fast negative going discharge ramp which would otherwise cause a glitch at the output. A sine wave can be shaped from the triangle with a CA3080 as shown. The output has about 2% THD when trimmed.

The variable width pulse output is derived by internally comparing the sawtooth to the pulse width input. The PWM pin is thus a voltage input with a fixed 10%/V sensitivity (100% at 0 Volts decreasing to 0% at 10V). The output is -1.4V low and 6.2V high.

HARD SYNC

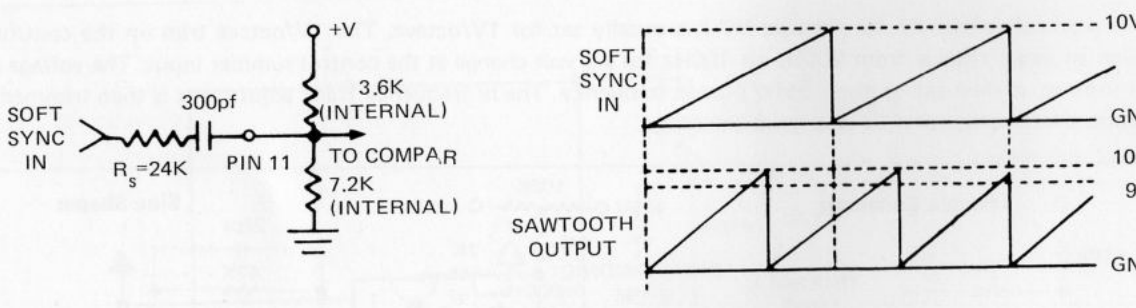
The hard sync input senses a falling edge, (such as another 2030's sawtooth discharge), and forces an immediate discharge of the synched 2030.



Hard Sync

SOFT SYNC

The soft sync input also accepts a falling edge but it will force discharge only if the synched 2030 is within $[240K/(R_s + 2.4K)]$ % of discharge. This enables one to phase lock two oscillators to frequencies that are exact small integer ratios of one another.



Soft Sync

Solid State Micro Technology for Music

SSM 2040

VOLTAGE CONTROLLED FILTER CIRCUIT

DESCRIPTION

The SSM 2040* is a four section filter whose cutoff frequency can be exponentially voltage controlled over a 10,000 to 1 range. This flexible building-block can be used in virtually any active filter design including lowpass, highpass, bandpass and notch. Rolloff characteristics can be selected to be Butterworth, Bessel, Chevyshev, Cauer or any other filter type.

Applications include tracking filters, organs, music synthesizers, music phase shifters and sound effects generation.

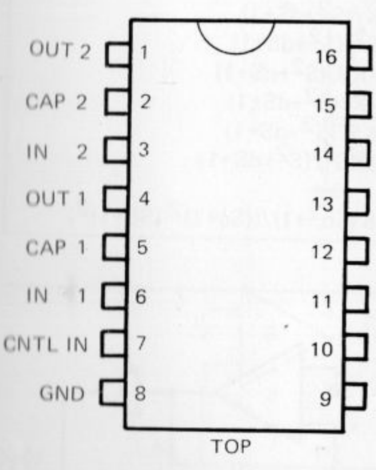
FEATURES

- +/- 15V Supplies
- Exponential Frequency Control Response
- 4 Filter Sections in One Package
- Low Noise
- Low Distortion
- Guaranteed Control Rejection Characteristics
- 10,000:1 Sweep Range

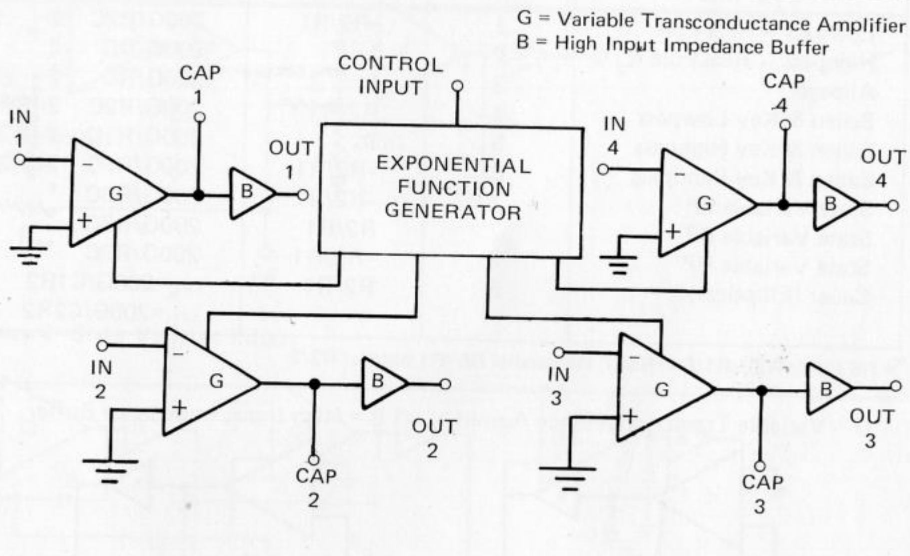
*Covered by U.S. Patent Number 3,969,682.

APPLICATIONS

- Voltage Controlled Filters:
  - Lowpass
  - Bandpass
  - Highpass
  - Allpass
  - Notch
  - Biquad
  - State Variable
  - Sallen & Key
  - Cauer
- Parametric Equalizers
- Music Synthesizers
- Music Phase Shifters
- Tracking Filters
- Low Distortion Sine VCO's



Pin Diagram



Block Diagram

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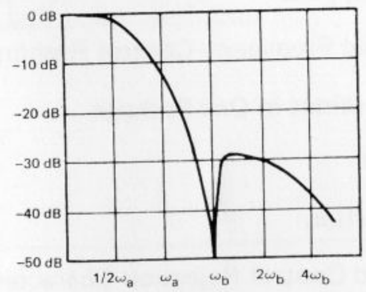
SPECIFICATIONS:

V_S = +/-15V, T_A = 25°C

SPECIFICATION	CONDITIONS	MIN	TYP	MAX	UNIT
Functional Range		10,000:1			
Input Offset, each cell			2	5	mV
Δ Input Offset, 4 cells in series	Vcntl= 0 mV, -90 mV Vcntl= 0 mV, +90 mV		0.6 0.6	3 3	mV mV
Transconductance	Vcntl= 0	1/10K	1/5K	1/3K	mhos
Equiv. Input Noise, each cell	20 Hz-20 KHz, Vcntl=-90 mV		0.5		uV RMS
Distortion (THD), Ein=30 mVpp	F=1 KHz, Vcntl=-90 mV		0.1		%
Tempco of Transconductance	Vcntl=0		+0.5		%/°C
Control Sensitivity			-18		mV/oct
Tempco of Control Sensitivity			0.33		%/°C
Power Supply Current	Vcntl=0	2	4	7	mA
Buffer Slew Rate			2		V/usec
Buffer Output Sink Current		425	560	750	μA

ABSOLUTE MAXIMUM RATINGS:

Any Pin to V-	36V
Current at any pin	20mA
Operating Temperature	0-70° C
Storage Temperature	-55-125° C
Power Dissipation	625 mW



CAUER FILTER RESPONSE – $\omega_b = 2\omega_a$

APPLICATIONS FIGURES:

Filter Type	Figure	Gain (K)	ω_o	d=1/Q	XFER char. $(S=j\omega/\omega_o)$
Lowpass – Real Pole	1	-R2/R1	200G/R2C	2	-K/(S+1)
Highpass – Real Pole	2	1	200G/RC	2	S/(S+1)
Allpass	3	1	200G/RC	2	(S-1)/(S+1)
Sallen & Key Lowpass	4	R2/R1	200G/R2C	2-(R4/R3)	$K/(S^2+dS+1)$
Sallen & Key Highpass	5	1	200G/R1C	2-(R3/R2)	$S^2/(S^2+dS+1)$
Sallen & Key Bandpass	6	-R2/R1	200G/R2C	2-(R2/R3)	$-KS/(S^2+dS+1)$
State Variable LP	7	-R2/R1	200G/R3C	*	$-K/(S^2+dS+1)$
State Variable BP	7	R2/R1	200G/R3C	*	$KS/(S^2+dS+1)$
State Variable HP	7	-R2/R1	200G/R3C	*	$-KS^2/(S^2+dS+1)$
Cauer (Elliptical)	8	R2/R1	ω_a=200G/C1R2 ω_b=200G/C2R2		$K(Sb^2+1)/((Sa+1)^2\ (Sb+1)^2)$

*= R5 (2R1+R2)/(R1(R4+R5)); R4 parallel R5=R1 parallel R2/2

G = Variable Transconductance Amplifier B = High Input Impedance Buffer.

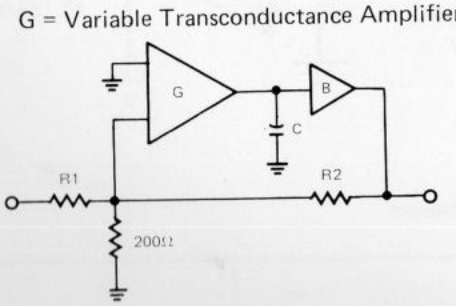


Figure 1–Lowpass Real Pole

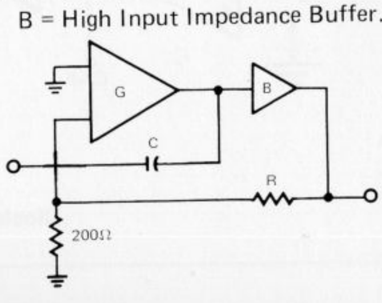


Figure 2–Highpass Real Pole

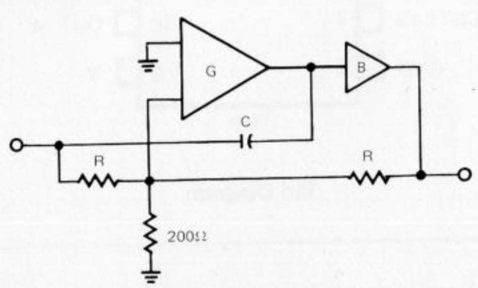


Figure 3–Allpass (Phase-shift)

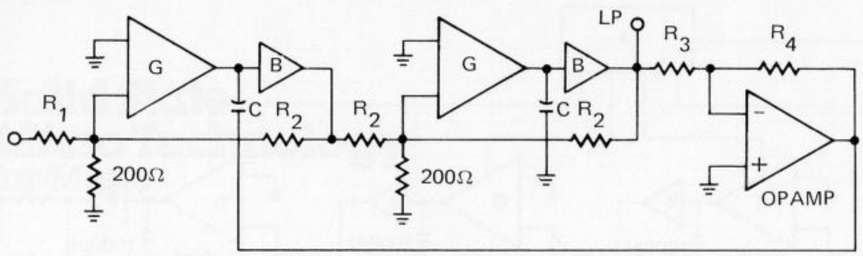


Figure 4—Sallen & Key Lowpass

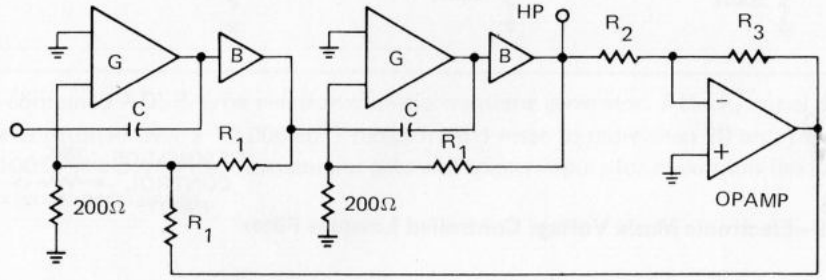


Figure 5—Sallen & Key Highpass

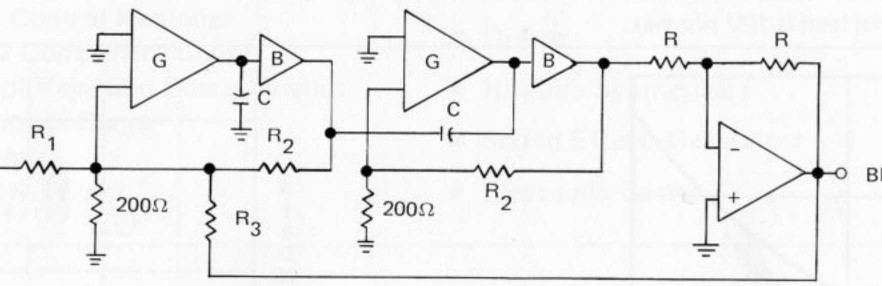


Figure 6—Sallen & Key Bandpass

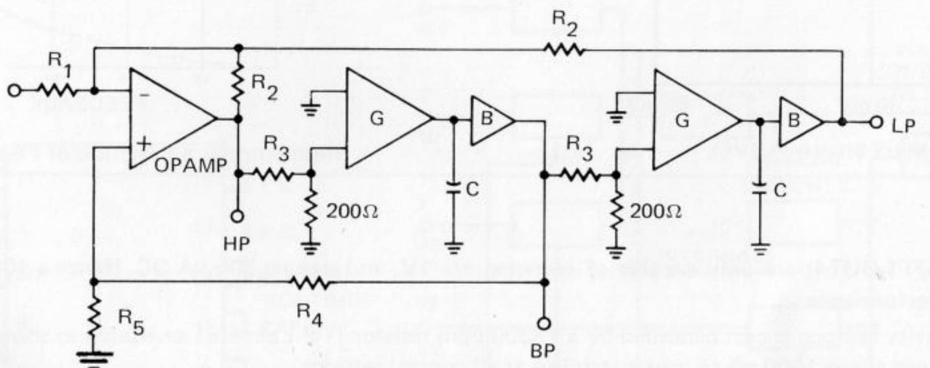


Figure 7—State Variable Filter

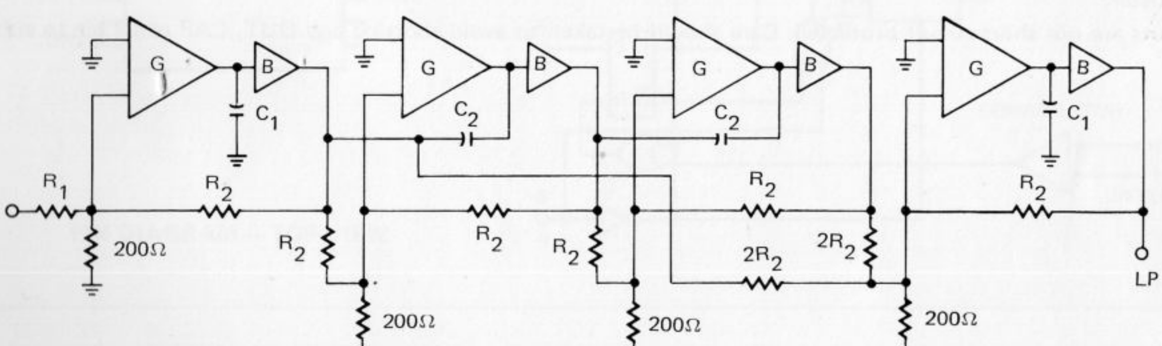


Figure 8—Cauer (Elliptical) Filter

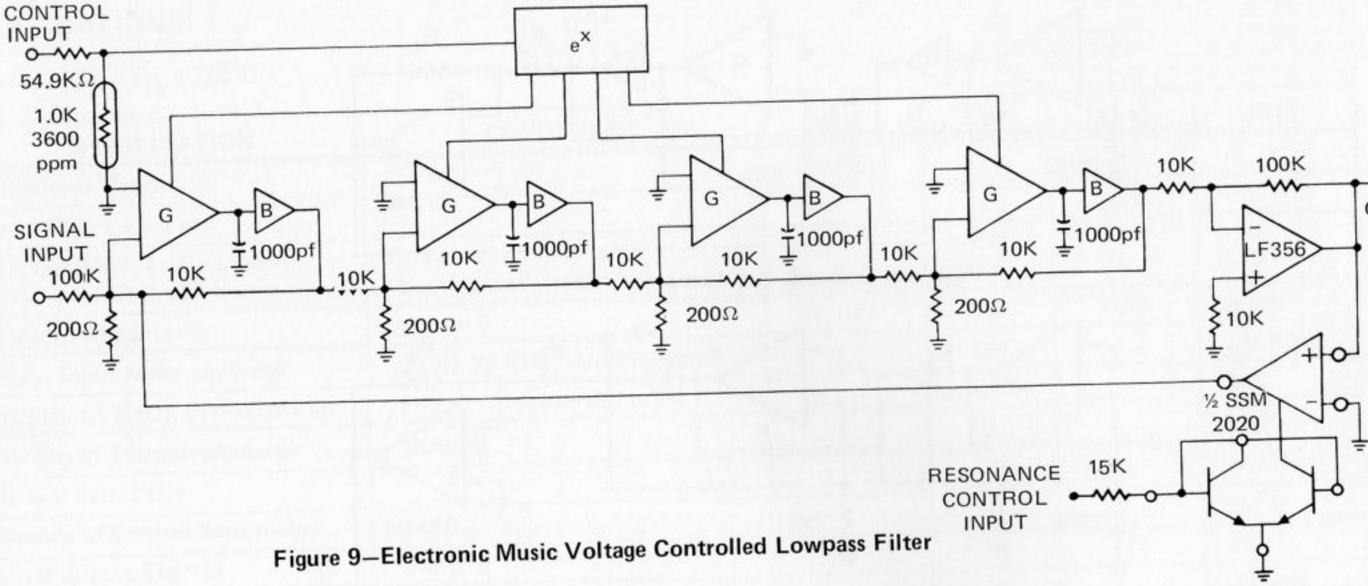
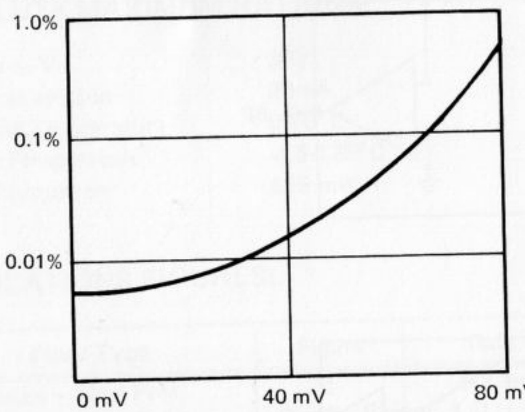
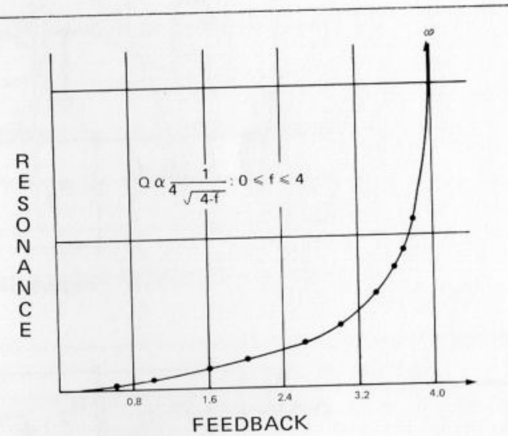


Figure 9—Electronic Music Voltage Controlled Lowpass Filter

Figure 9 is a voltage controlled lowpass filter with voltage controlled resonance for electronic music applications. The frequency control input sensitivity is 1 volt/octave, temperature compensated. At high resonance settings the filter will oscillate with a pure sine wave. The output signal level is 10V p/p max.



THD vs. INPUT SIGNAL LEVEL



Resonance as a Function of Feedback for Figure 9

DESIGN HINTS:

1. The output pins (OUT1-OUT4) are only capable of swinging +/- 1V, and sinking 500 μA DC. Hence a 10K feedback resistor & load will give good performance.
2. Control input sensitivity tempco is best cancelled by a +3300 ppm resistor (Tel Labs Q81 or equiv.) as shown in Figure 9.
3. C values should be kept above 1000 pF to insure stability at all control settings.
4. The 200 ohm attenuating resistor is chosen for optimal control rejection. Other values can be used with some degradation of this parameter.
5. The outputs are not short-circuit protected. Care should be taken to avoid shorting any OUT, CAP or IN pin to either supply.

Solid State Micro Technology for Music

SSM 2050

VOLTAGE CONTROLLED TRANSIENT GENERATOR

DESCRIPTION

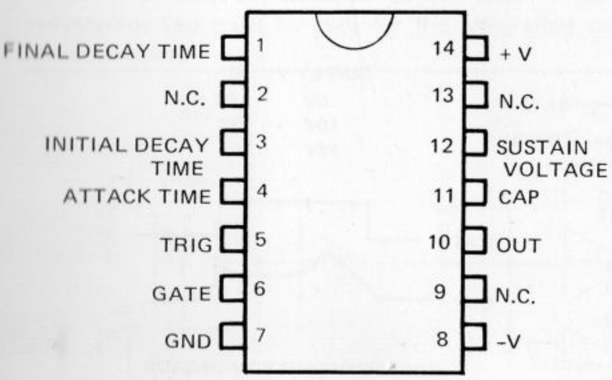
The SSM 2050 is a self-contained ADSR type electronic music transient generator. Attack, initial decay and final decay times can be exponentially voltage controlled over a 10,000 to 1 range from 1 msec to more than 10 sec. The sustain level is linearly voltage controllable from 0 to 100%. The device has independant gate and trigger inputs for maximum flexibility and much effort has been taken to minimize the external parts count.

FEATURES

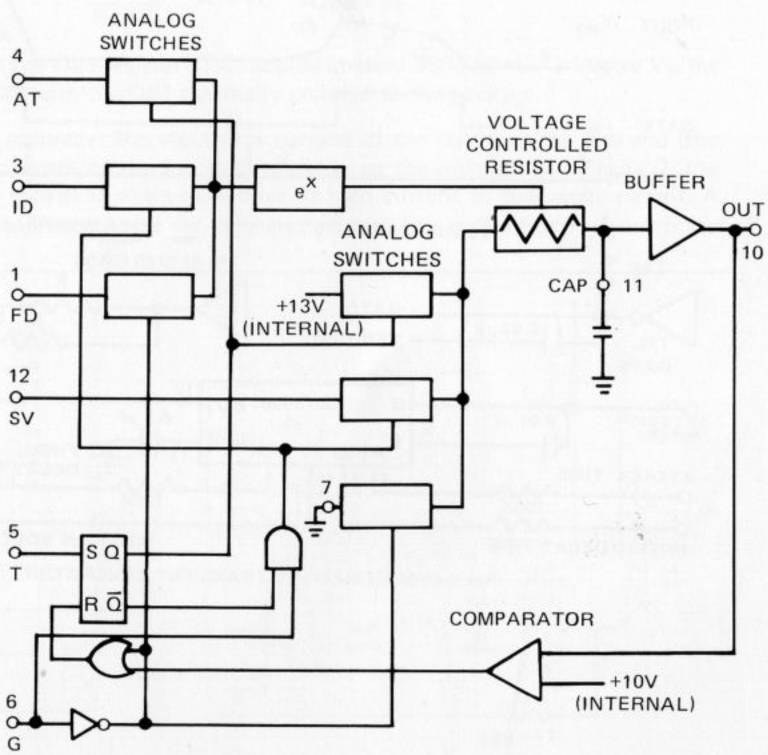
- +/- 15V Supplies
- Exponential Time Control Response
- Minimum External Component Count
- Guaranteed Control Rejection Characteristics
- 10,000:1 Time Control Range
- Full ADSR Response
- Independent Gate & Trigger

APPLICATIONS

- Music Synthesizers
- Organs
- Rhythm Synthesizers
- Sound Effects Generators
- Electronic Games



PIN DIAGRAM – TOP VIEW



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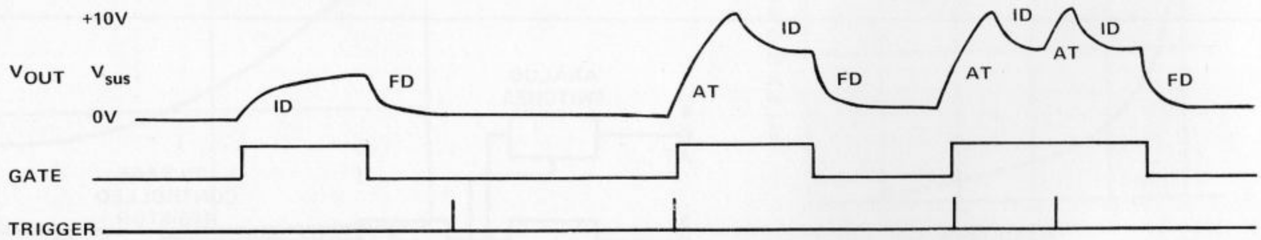
SPECIFICATIONS:
V_S = +/-15V, T_A = 25°C, CAP = 0.1 μF

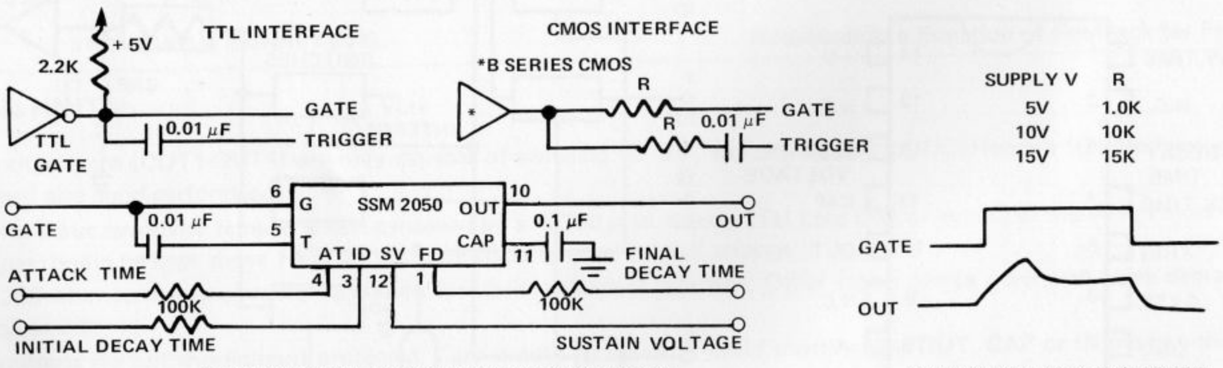
OPERATING TEMPERATURE
0 to 75°C

STORAGE TEMPERATURE
-55 to +125°C

PARAMETER	MIN	TYP	MAX	UNIT
Time Range	2-20000	1-100000		msec
Offset, Gate=off	±250	±50		mV
Time Constants V_{IN} = 0	50	100	200	msec
Offset V_{SUS}-V_{OUT} Gate = ON	-1	0	+1	V
Gate & Trig On Voltage Current V_{IN} = 1.5V		1.0 500	1.5 750	V μA
Output Noise		0.5		mV RMS
V_{attack}	10	10.5	11	V
Final Decay Control Rejection V_{OUT}; V_{cntl} = 0 –+120mV V_{cntl} = 0 –+120mV		30 30	150 150	mV mV
Control Input Impedance	2.3	3.1	3.9	kohm
Time Control Sensitivity		+18		mV/octave

AT, ID, FD indicate times controlled by Attack, Initial Decay and Final Decay Time Control Inputs respectively. (A positive voltage increases the time constant.) All are nominally exponential approaches to +13V, Sustain Voltage and Ground respectively.





POLY SYNTHESIZER VC TRANSIENT GENERATOR

TIME SENSITIVITY 2V/DECADE

Solid State Micro Technology for Music

SSM 2030, 2040, 2050 Application Note

APPLICATIONS FOR AUDIO AND ELECTRONIC MUSIC

By: Ron Dow, V.P. Engineering, SSMT
Dave Rossum, Chief Engineer, Eμ Systems

An Integrated Circuit VCO for Electronic Music

The most critical element in any voltage controlled synthesizer is the voltage controlled oscillator. Design requirements include high temperature stability, 1000 to 1 sweep range, multiple waveform outputs, and an exponential sweep control with excellent log conformity over the entire audio range. The Solid State Music 2030 is the first VCO in Integrated Circuit form to address the needs of electronic music systems.

The circuit consists of a pair of matched logging transistors, a precision current mirror, a fast ultra-low leakage buffer, wave-shaping circuits for pulse and triangle outputs, a fast comparator and a discharge circuit, which includes a capacitorless one-shot. (Figure 1) In addition, provision has been made to allow linear FM and synchronizing the oscillator to the output of another.

The frequency control circuit shown in Figure 2 is similar to many modular designs now in use. A low input bias op amp is used to force the current in Q_1, to be equal to the reference current established by R_1 and the linear FM voltage (if any). The current in the output transistor Q_2 is:

$$I_o = (V_+/R_1 + V_L/R_2)\ e^{-V_b q/KT}$$

As one can see, the term in the exponent is temperature dependent. This problem can be addressed by making V_b temperature dependent.

$$\frac{d}{dT}\frac{KT}{q} = 3300\text{ppm/C}^\circ\ @\ 25^\circ\text{C} \qquad V_b = \frac{V_e R_3}{R_3 + R_4} \qquad R_3 = 1\text{K} \quad R_4 = 54.9\text{K}$$

Since R_4 is large compared to R_3 one can use a Tel Lab type Q81 resistor (T.C. approximately 3600 ppm/c^0) to give V_b the necessary temperature dependence. Best results are obtained with the Q81 thermally coupled to the package.

Two other sources of error exist which can affect sweep accuracy; the input bias current of the buffer at the low end (the 2030 specification is I_b<1nA), and the bulk emitter resistance of the logging transistors at the high end. In Figure 2, the diode-R_5 network compensates for this latter effect by providing extra base drive at high current to the sweep circuit. A polystyrene cap must be used for the integrating capacitor because of its temperature stability and dielectric absorption characteristics.

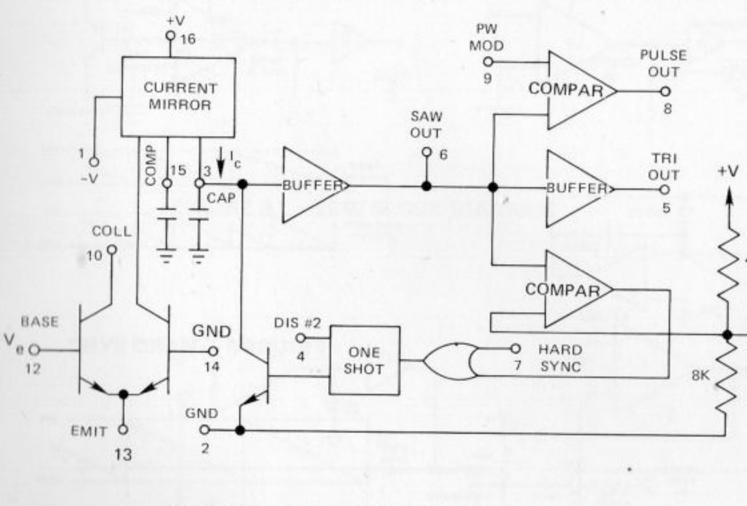


FIGURE 1 – BLOCK DIAGRAM

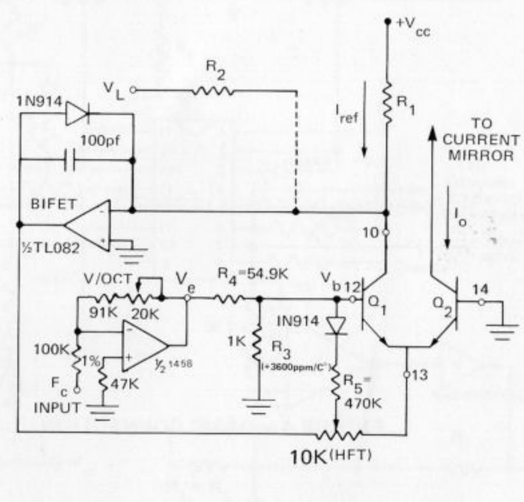


FIGURE 2 – CONTROL CIRCUIT

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For proper operation the Volt/octave trimmer is adjusted to give a true octave at 200Hz to 400Hz. The HFT pot is then trimmed for a true octave at 4000Hz to 8000Hz. Using this procedure one can get better than 0.25% absolute sweep accuracy from 20Hz to 10kHz with 0.1% in the critical range between 100Hz to 8000Hz. Since the small remaining errors tend to be similar from oscillator to oscillator, matching between oscillators is even better than the absolute frequency accuracy; an important benefit in designing polyphonic systems.

Within the chip, the exponential current is mirrored, and direct integration yields a sawtooth whose discharge results from the comparator triggering the one-shot. The sawtooth waveform is 0-10V.

The sawtooth and "triangle" output pins are both emitter followers from the internal sawtooth. The triangle waveshaping circuit (Figure 3) uses the sawtooth output and an emitter follower from the same point which is biased to give half a sawtooth. The subtraction performed by the 1458 of the sawtooth from this waveform results in a triangle output. A true 741 type op amp must be used here, as its lower slew rate ignores the fast negative going discharge ramp which would otherwise cause a glitch at the output. A sine wave can be shaped from the triangle with a CA3080 as shown. The output has about 2% THD when trimmed. (Note 1)

The variable width pulse output is derived by internally comparing the sawtooth to the pulse width input. The PWM pin is thus a voltage input with a fixed 10%/V sensitivity (100% at 0 Volts decreasing to 0% at 10V). The output is -1.4V low and 6.2V high.

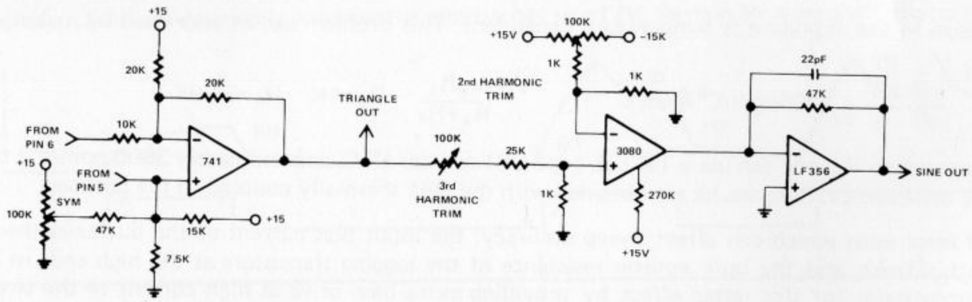
The hard sync input senses a falling edge, (such as another 2030's sawtooth discharge), and forces an immediate discharge of the synched 2030. The soft sync input also accepts a falling edge but it will force discharge only if the synched 2030 is within $[240K/(R_S + 2.4K)]$ % of discharge. (Note 2) This enables one to phase lock two oscillators to frequencies that are exact small integer ratios of one another. (Figures 5, 6)

The auxiliary discharge output can be used to drive a large geometry transistor for discharging capacitors from 2000pF to 10,000pF in LFO applications (oscillators used for modulation purposes).

Additional important specs are ±50ppm/C temperature stability of the basic oscillator loop (exponential input grounded), and a 1,000,000 to 1 sweep range from 0.1Hz to 100KHz with a 1000pF integrating capacitor.

(1) Bernie Hutchins, *Musical Engineer's Handbook,* Chapter 50.

(2) David P. Rossum, *Electronotes,* Number 51, page 10.



TRIANGLE CONVERTER

SINE SHAPER

FIGURE 3

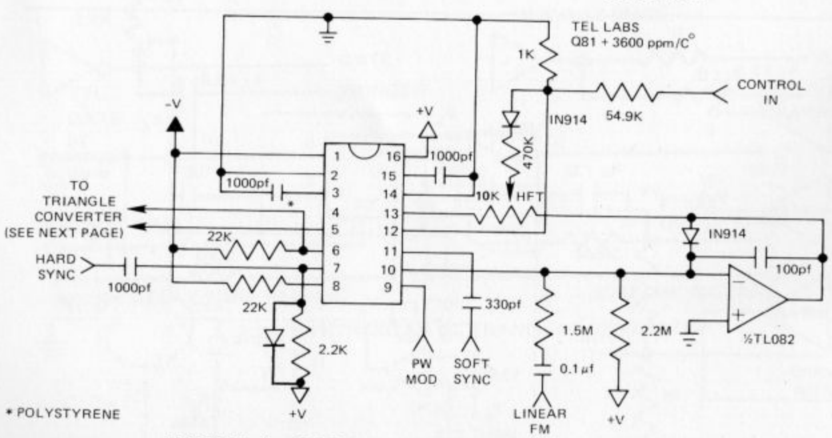


FIGURE 4 – BASIC CONNECTION

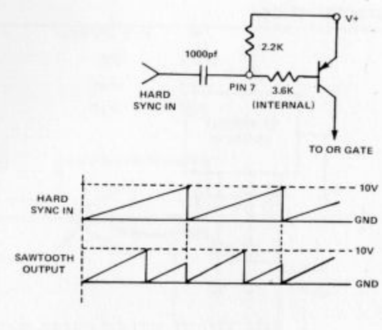


FIGURE 5 – HARD SYNC

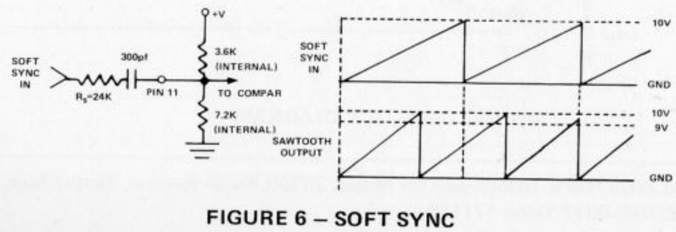


FIGURE 6 – SOFT SYNC

2040 VOLTAGE CONTROLLED FILTER

The 2040 Voltage Controlled Filter contains four identical filter stages, all of which are simultaneously controlled by the same exponential function generator. The block diagram of the filter is given in Figure A1. The equation describing the variable transconductance amplifier is:

$$I_o \cong (V_+ - V_-)\, G \quad ; \quad G = I_c \,/\, 52\text{ mV @ }25°\text{C}$$

The generalized voltage controlled filter connection is shown in Figure A2. The differential form for the equation describing the filter operation is:

$$\frac{d(V_o - V_{ib})}{dt} = \frac{-G\, R_a\, (R_f V_{ia} + R_i V_o)}{C\,(R_i R_f + R_a R_i + R_a R_f)}$$

Solving this equation for the general filter function we find:

$$V_o = \frac{V_{ib}(S/\omega_o) - K\, V_{ia}}{1 + (S/\omega_o)}$$

$$\text{where } S = j\omega\ ,\ \omega_o = \frac{G\, R_a R_i}{C(R_a R_i + R_a R_f + R_i R_f)}\ , \quad K = R_f / R_i$$

From the above general equation, we can derive the three specific single stage cases. Figure A3-A5 show the lowpass, highpass, and allpass (phase shift) connections for the single stage, and their equations are:

LOWPASS REAL POLE: $V_o = -K\, V_i \,/\, (1 + S/\omega_o)$

HIGHPASS REAL POLE: $V_o = V_i (S/\omega_o) \,/\, (1 + S/\omega_o)$

ALLPASS NETWORK: $V_o = V_i (S/\omega_o - 1) \,/\, (S/\omega_o + 1)$

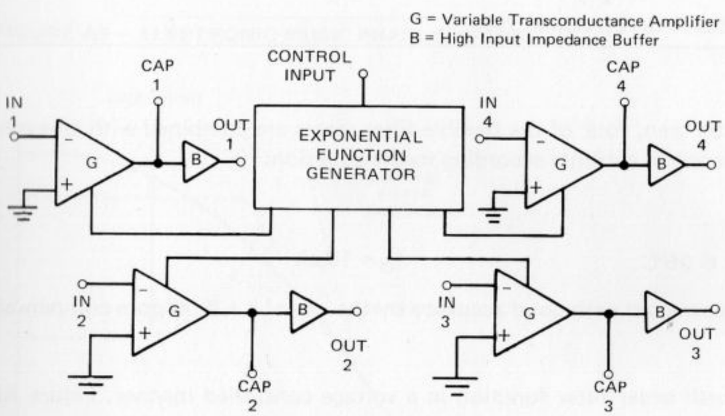


FIGURE A1 – 2040 BLOCK DIAGRAM

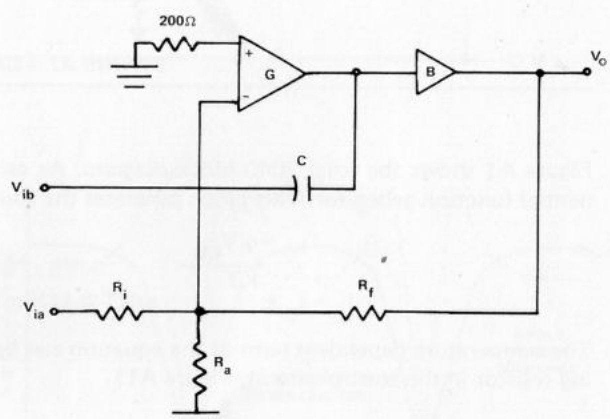


FIGURE A2 – GENERALIZED 2040 CONNECTION

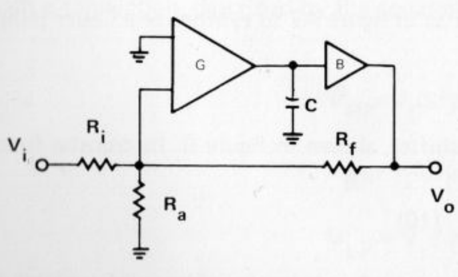


FIGURE A3 – LOWPASS REAL POLE

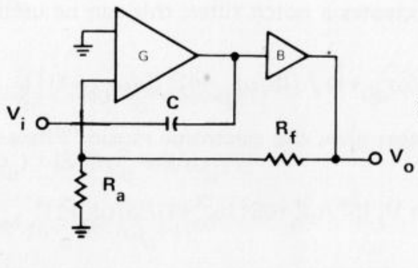


FIGURE A4 – HIGHPASS REAL POLE

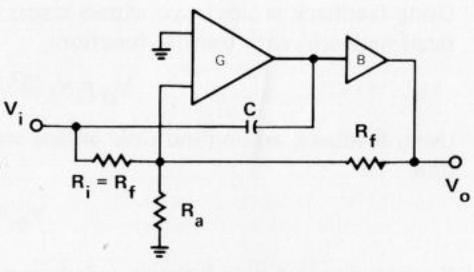


FIGURE A5 – ALLPASS

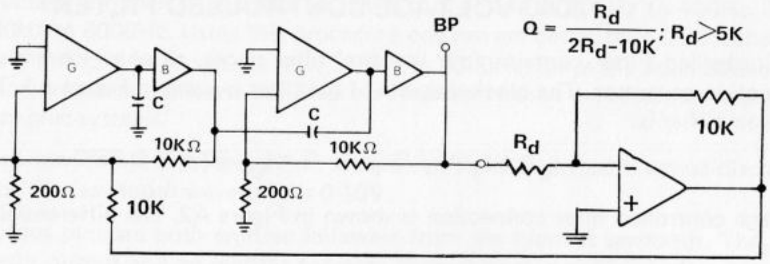
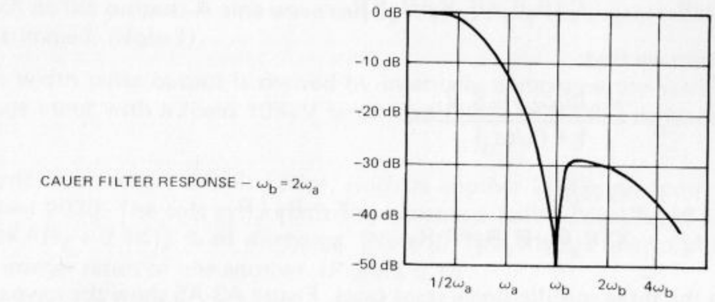


FIGURE A6 – SALLEN & KEY BANDPASS FILTER



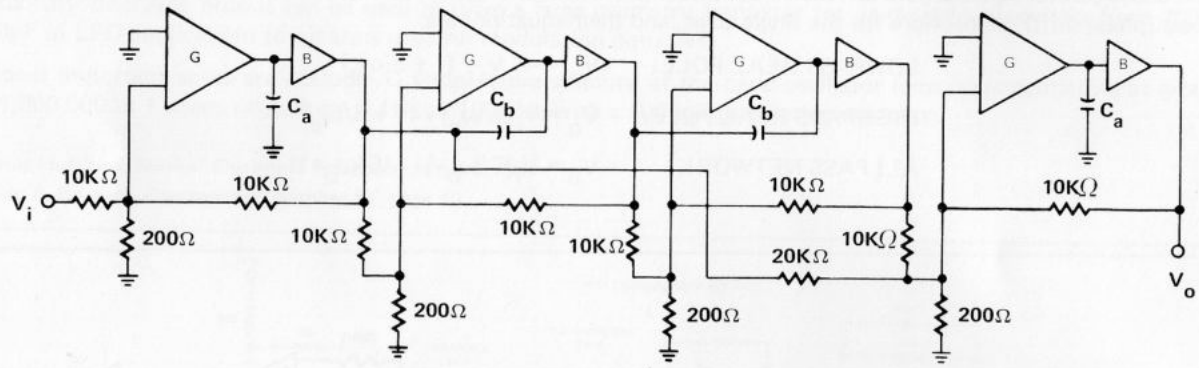


FIGURE A7 – CAUER FILTER USING 2040

Figure A1 shows the total 2040 block diagram. As can be seen, four of the flexible filter stages are combined with an exponential function generator. This block generates the four control currents according to the equation:

$$I_c = I_o e^{\frac{-q V_{ctl}}{KT}} \quad ; \; q/KT \cong 1/26\text{mV} \text{ @ } 25^\circ\text{C} \qquad I_o = 10\mu A$$

The temperature dependent term of the equation can be corrected with good accuracy by the use of a +3600 ppm compensating resistor in the control circuit, Figure A11.

The result is a circuit capable of implementing any fourth order filter function in a voltage controlled manner. Figure A8 shows the classical "four-pole" electronic music lowpass filter.

Using additional feedback allows one to synthesize pole pairs with resonance. Figure A6 shows a Sallen & Key bandpass filter, with transfer function:

$$V_o = -V_i\,(S/\omega_o)\;/\;(S^2/\omega_o^2 + (2 - R_d/10K\Omega)(S/\omega_o) + 1) \qquad (8)$$

Using feedback around two allpass stages creates a notch filter; this can be used as in figure A7 to synthesize a Cauer (elliptical) network, with transfer function:

$$V_o = V_i\,(S^2/\omega_{ob}^2 + 1)\;/\;[(S/\omega_{oa} + 1)^2 (S/\omega_{ob} + 1)^2] \qquad (9)$$

Using feedback around multiple allpass stages gives the electronic music "Phase-shifter, shown in figure 9. Its transfer function:

$$V_o = V_i\,(S^4/\omega_o^4 + 6S^2/\omega_o^2 + 1)/(S/\omega_o + 1)^4 \qquad (10)$$

By including a 4-pole 2-throw switch and a few extra components, one can implement a 4-pole lowpass filter with variable Q and a 4-pole phasor that has variable regeneration with a single 2040 (Figure A10).

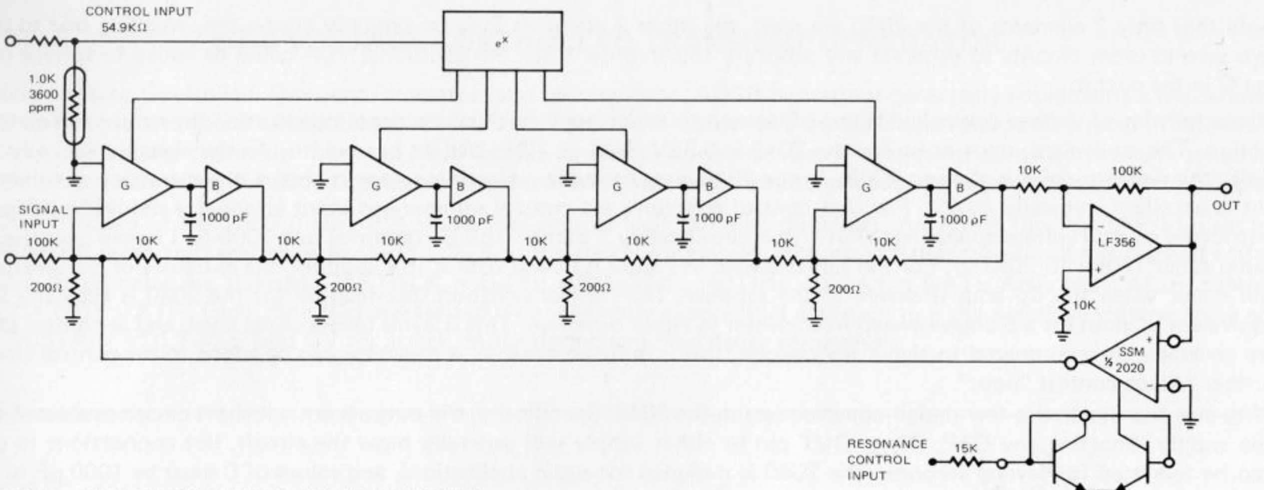


FIGURE A8 – ELECTRONIC MUSIC LOWPASS FILTER

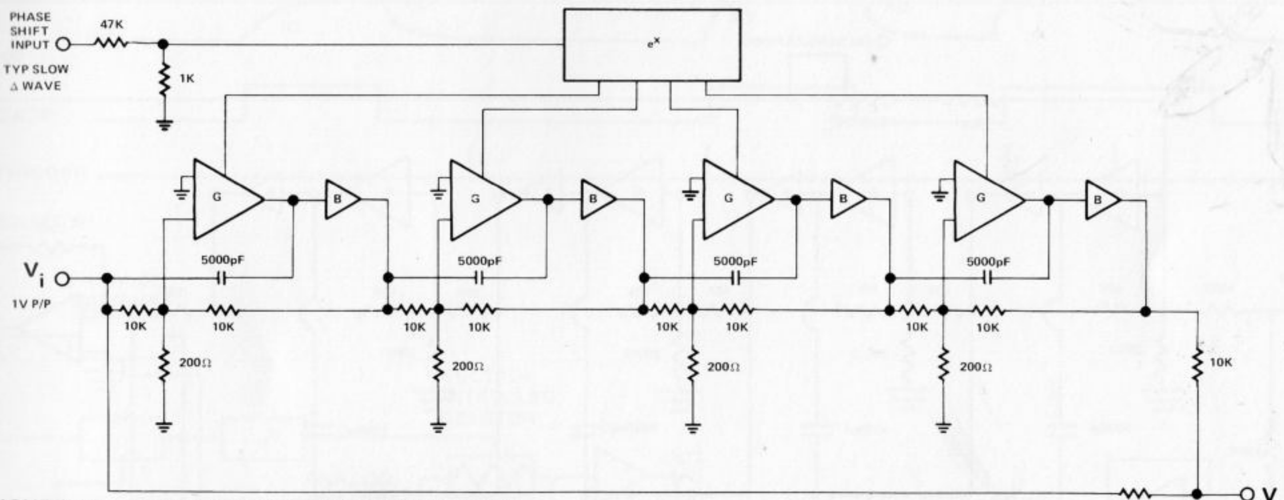
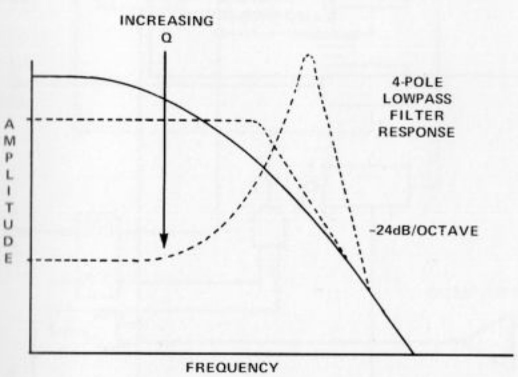
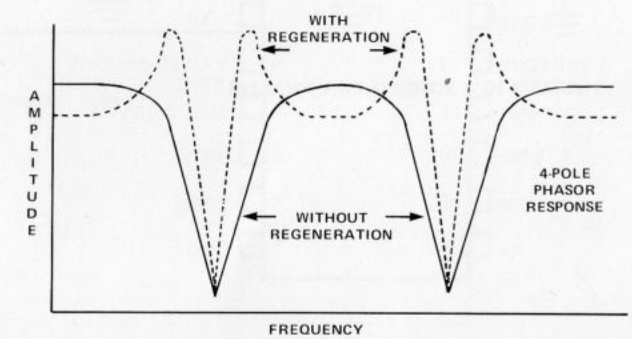


FIGURE A9 – ELECTRONIC MUSIC PHASE SHIFTER





It is also possible to use the elements of the 2040 in the state variable configuration to synthesize filters. Figure A12 shows such a connection, described by the equations:

$$V_{HP} = V_i (S^2/\omega_o^2) / (S^2/\omega_o^2 + dS/\omega_o + 1) \qquad (11)$$

$$V_{BP} = V_i (S/\omega_o) / (S^2/\omega_o^2 + dS/\omega_o + 1)$$

$$V_{LP} = V_i / (S^2/\omega_o^2 + dS/\omega_o + 1)$$

$$d = 3R_1 / (R_1 + R_2)$$

Note that only 2 elements of the 2040 are used, the other 2 elements may be similarly connected, enabling one to use the two second order circuits to generate any arbitrary fourth order filter. An additional VCA could be added to voltage control the Q in the circuit.

Characteristics of voltage controlled filters of interest to design engineers include signal/noise ratio, distortion, and control rejection. The equivalent input noise for the 2040 is 0.5μV RMS at 20Hz-20KHz bandwidth. As the signal is typically 20mV RMS, the signal-to-noise is close to 90dB. As the 2040 is operating in a closed loop environment, the distortion at full level input is excellent, typically 0.02%. For best control rejection, the control summer and input attenuator should be designed to restrict the control voltage appearing at Pin 7 to approximately ± 90mV. This corresponds to a 1000-to-1 sweep over the entire audio range (20Hz to 20KHz). For the values shown in Figure A11 and with ± 15V supplier, the extremes of the sweep range will occur when the op amp is driven to the supplier. The control rejection specification for the 2040 is typically 0.6mV equivalent at input for a 5-octave sweep from center in either direction. This is 37dB below signal level, and as control changes are generally slow compared to signal frequencies, this is quite acceptable. A capacitor can be added to the control circuit to further reduce control "pop."

Chip size has dictated a few design compromises in the 2040. Specifically, the outputs are not short circuit protected against the supply. Shorting any CAP, IN, or OUT pin to either supply will generally blow the circuit, but connections to ground can be tolerated for several seconds. The 2040 is designed for audio applications, and values of C must be 1000 pF or higher to insure stability. The transconductance amplifiers are not internally temperature compensated, and have a temperature coefficient of nominally 0.5%/degree C. This can be compensated by a temperature dependent Ra.

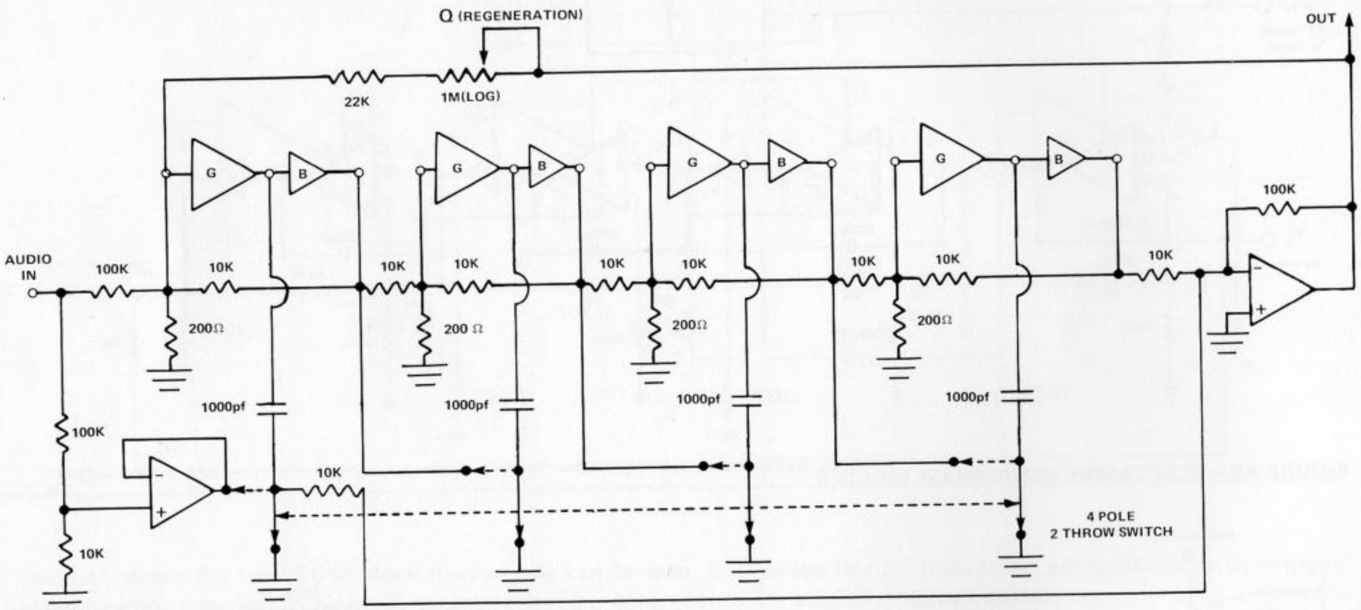


FIGURE A10 – 2040 FILTER/PHASOR

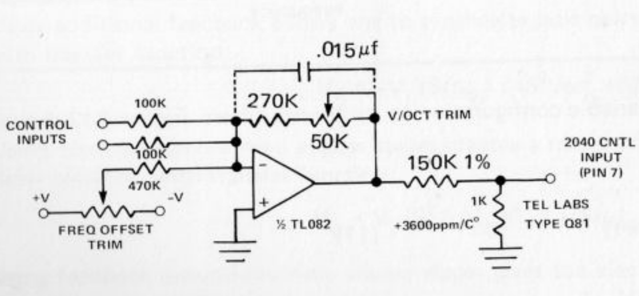


FIGURE A11 – 2040 PRECISION CONTROL SUMMER FOR ±15 SUPPLIES

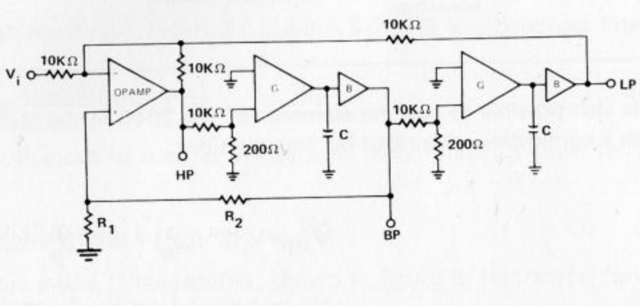


FIGURE A12 – STATE VARIABLE FILTER USING 2040

2050 VOLTAGE CONTROLLED TRANSIENT GENERATOR

The 2050 Voltage Controlled Transient Generator (also call envelope, ADSR or contour generator) implements a well-known electronic music function in a voltage controlled manner. The usual function of a transient generator is illustrated in figure B1. Classically, the transient generator has been implemented by charging and discharging a capacitor with electronically switched panel controls. In contrast, the 2050 uses a voltage controlled resistor internally to generate the nominally exponential slopes.

The block diagram of the 2050 (figure B2) shows the internal logic defining the states. An attack flip/flop (AF/F) is set by the TRIGGER pulse and reset by either NOT GATE or the attack comparator determining that the output has reached +10V. Then the three states are defined: ATTACK = GATE AND AF/F, INITIAL DECAY = GATE AND NOT AF/F, FINAL DECAY = NOT GATE. Each state is characterized by a nominally exponential approach to a characteristic voltage; these are +13V, SUSTAIN VOLTAGE, and OV for attack, initial decay, and final decay respectively.

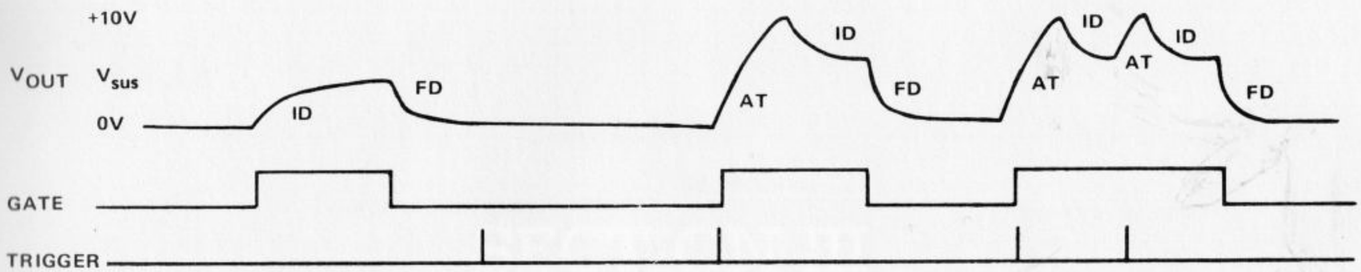
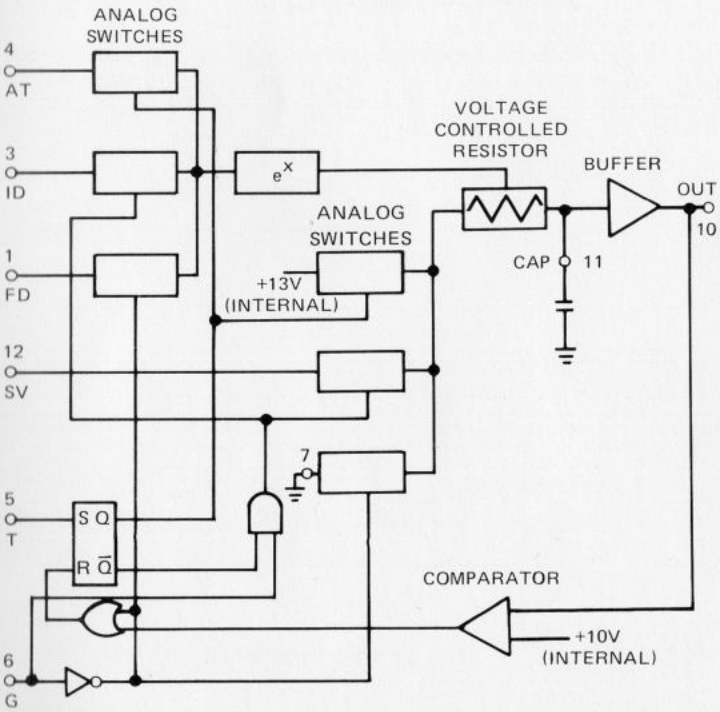


FIGURE B1



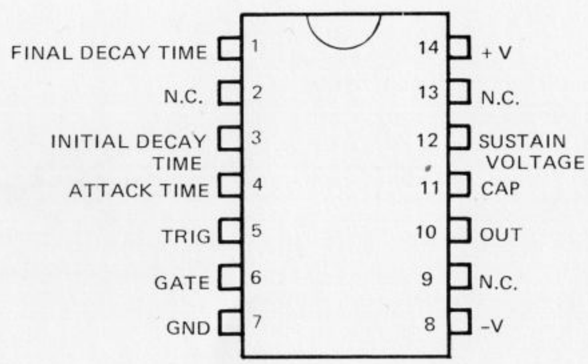


FIGURE B2

2050 PIN OUT TOP VIEW

The input stage of the 2050 logic inputs is given in Figure B3. The lateral PNP structure protects the logic inputs against all voltages. The input sensitivity specification for the logic inputs is 750μa or 1.5V max. This is the minimum current and voltage required to trigger the 2050. The optimum logic interfaces to meet these requirements are given in Figure B4.

As with the 2040, a few design compromises had to be made to integrate the voltage controlled transient generator. The voltage controlled resistor is not perfectly linear, hence the approaches are only nominally exponential. The deviation is less than +/-5%, though. Also, there is a control rejection specification determining the maximum possible change in the output voltage approached in final decay mode (ideally OV); this is typically 30 mV for a time constant change 2.5 decades either way from nominal.

The attack, initial decay, and final decay inputs of the 2050 have a sensitivity of 60 mV/decade, and an input impedance of 3100 ohms +/- 25%. This latter tolerance is due to the variation of chip resistors. The input impedance for the 3 inputs on any given chip is matched to about 1%, though. Hence in polyphonic systems, where many matched 2050's are to be used, the 2050's must be selected for similar control input impedance. Fortunately, this can be trivially measured with a low current ohmmeter as the resistance from pin 1 to pin 7. Additionally, the initial time constants on the chip can vary +/- 50%, and as the 0.1 μF cap is often of poor tolerance as well, a trim for initial time constant is useful. Again, the time constants within the chip are well matched, so a single trim, as shown in figure B5, will suffice. With the values shown below, the time sensitivities are 1V per octave.

The minimum output current spec is 750μA with a +15 supply. To ensure stability, the P.C. line from the output should be kept as short as possible.

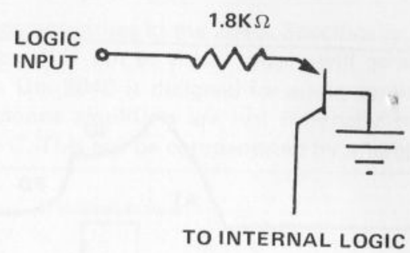


FIGURE B3

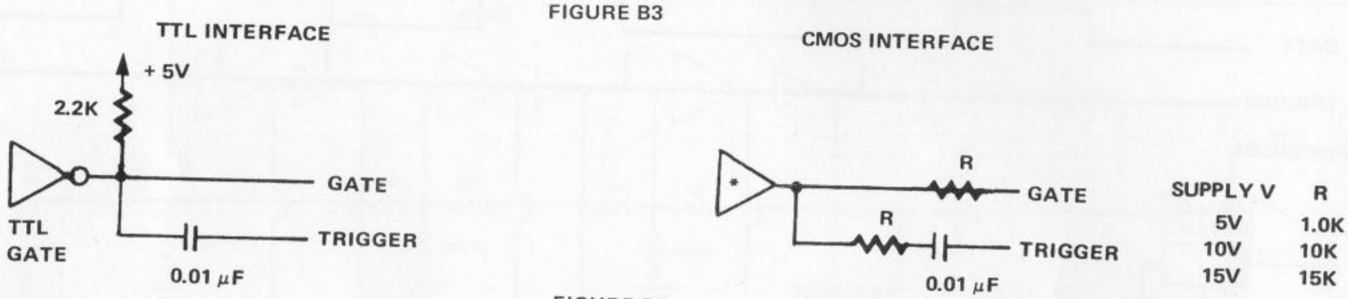


FIGURE B4

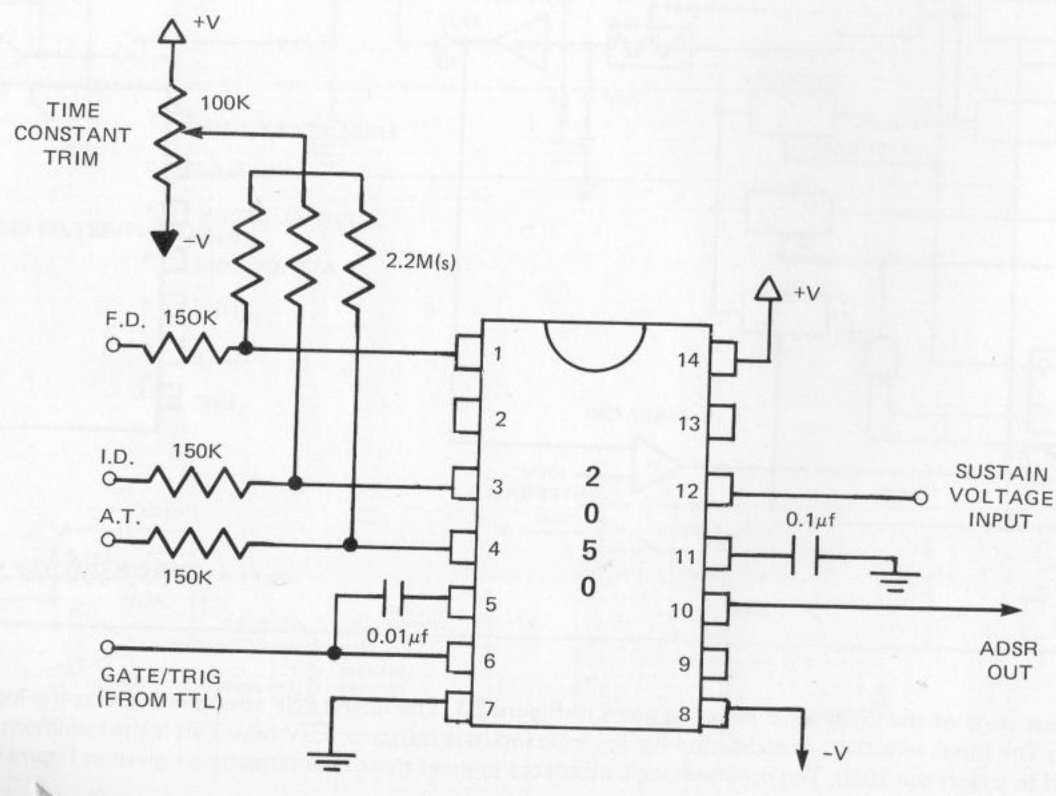


Figure B5—Typical Connection

SECTION III
NEW PRODUCTS

Solid State Micro Technology for Music

ADVANCE INFORMATION

SSM 2044

4-POLE VOLTAGE CONTROLLED FILTER

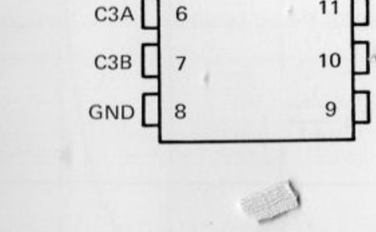
DESCRIPTION

The SSM2044 is a low cost 4-pole voltage controlled filter whose design has been optimized for use as an electronic music lowpass filter. On-chip voltage control of resonance allows direct and easy interfacing with programmers and controllers. A novel filtering technique* provides extended control range, low noise and high control rejection for "pop"-free performance. The filter can also be used as a low distortion sinewave oscillator. No external ladder network is required making the device a real cost and space saver in polyphonic applications.

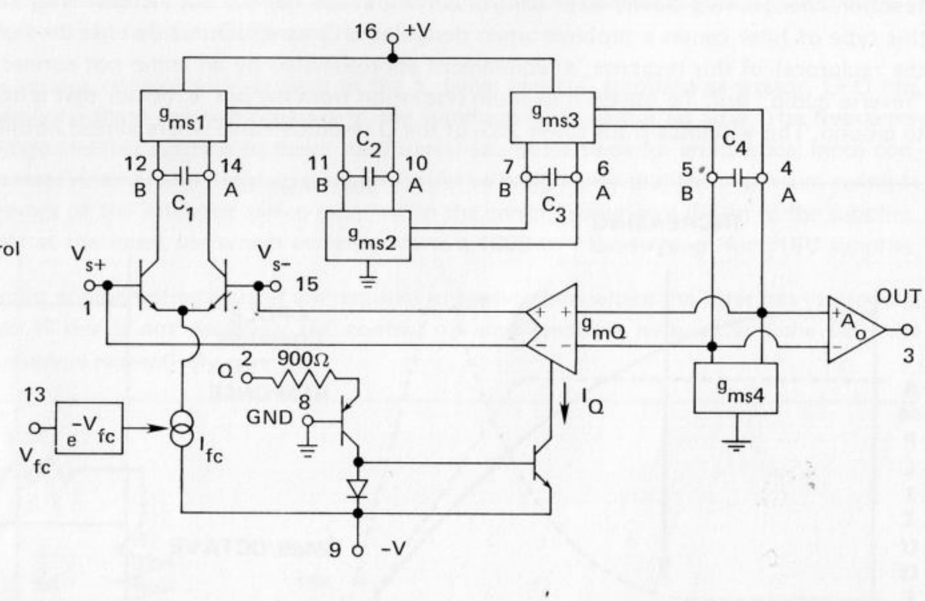
FEATURES

- Low Cost
- High Control Rejection (40db typical for 1000 to 1 sweep)
- ± 18V to ± 5V Supplies
- Minimum External Parts Count
- Current Output
- 90db Signal-to-Noise
- 10,000 to 1 Minimum Sweep Range
- On-Chip Resonance Control
- Differential Signal Inputs
- Stable Resonance Over Frequency Sweep

*Patent applied for.



Pin Out – Top View



Functional Block Diagram

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SPECIFICATIONS*

@ $V_s = \pm15$ and $T_A = 25°C$

STORAGE TEMPERATURE

-55°C to +125°C

OPERATING TEMPERATURE

0°C to +70°C

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
Positive Supply Range	+5	+15	+18	V	
Negative Supply Range	-5	−15	−18	V	
Positive Supply Current	1.0	1.4	2.0	mA	V_{FC} = GND
Negative Supply Current	4.5	6.2	8.0	mA	V_{FC} = GND
Frequency Control Range	10,000:1	50,000:1	—		
Frequency Control Feedthrough	—	-40db	-30db		$V_{s+} = V_{s-}$ = GND $-90mV \leqslant V_{FC} \leqslant +90mV$
Output Offset $I_O/I_{O Max}$		0.05	0.2		$V_{s+} = V_{s-} = V_{FC}$ = GND
Frequency Control Offset f/fnom	0.6	1	1.5		Untrimmed
Q Control Input Impedance	675	900	1200	Ω	$V_{QC} \geqslant 0.7V$
Q Current at Oscillation	400	425	450	μA	$-90mV \leqslant V_{fc} \leqslant +90mV$
Q Control Feedthrough	—	-30db	-20db		Untrimmed
Q Control Feedthrough			-60db		Trimmed
Max Available Control Current	1.25	1.7	2.2	mA	
Freq. Control Input Range	-120	—	+180	mV	
Max Output Signal Current $I_{O Max}$	± 300	± 400	± 520	μA	
Signal to Noise	—	90db	—		

*FINAL SPECIFICATIONS MAY BE SUBJECT TO CHANGE.

Figure 1 below shows the amplitude vs frequency response for the 2044 at different Q or resonance settings. The solid curve is the response of the filter at minimum Q which is a gradual rolloff approaching -24db/octave at high frequencies. As Q is increased, low frequency components are suppressed and components near the cutoff frequency are emphasized. For all Q settings below oscillation the final rolloff at high frequencies is -24db/octave. At high Q settings the filter will oscillate with a pure sinewave at the cutoff frequency. This waveform can be used as a tone source if the design procedures given below are followed.

The second figure below shows Q or resonance of a four-pole lowpass filter as a function of feedback or Q control current. The function changes very slowly with control current at the low end but increases very rapidly as oscillation is approached. In general, this type of filter causes a problem when designing a Q panel control that has the right *feel*. The optimum control pot would have the reciprocal of this response; a requirement approximated by an audio pot connected in reverse of its normal configuration: a "reverse audio" pot. To obtain maximum resolution from the pot, a resistor that is one-third of its value can be connected in series to ground. This will discard the lower 25% of the Q response curve where almost nothing happens. Figure 4.

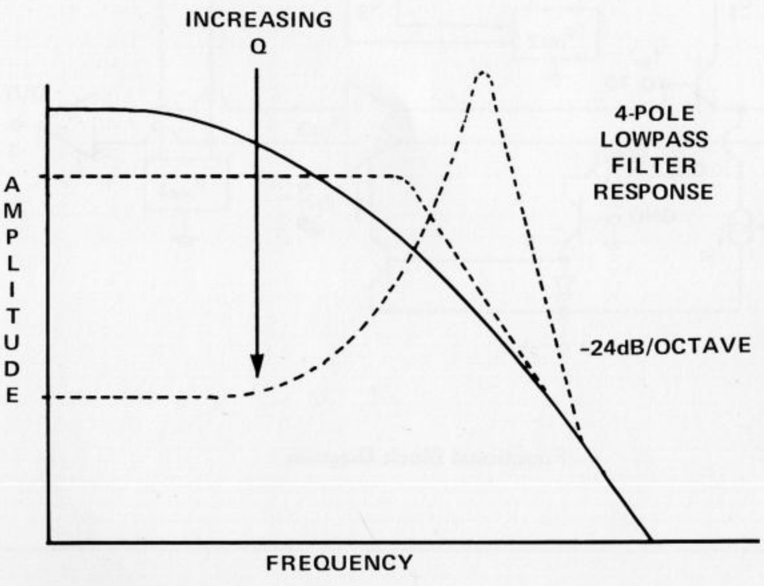


Figure 1

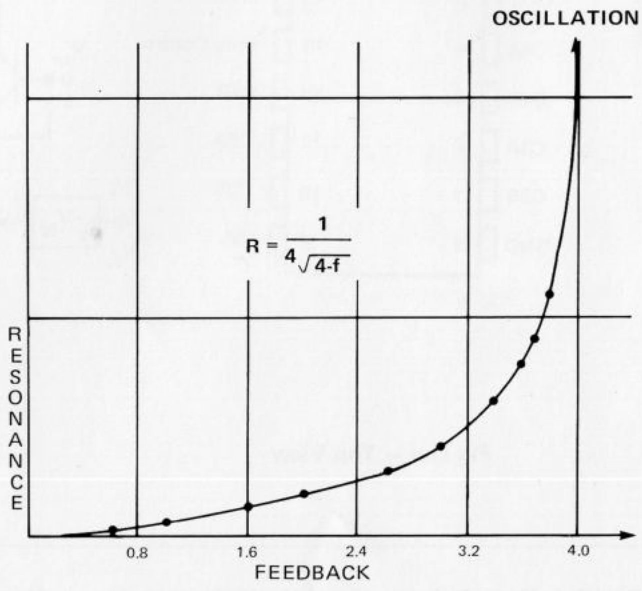
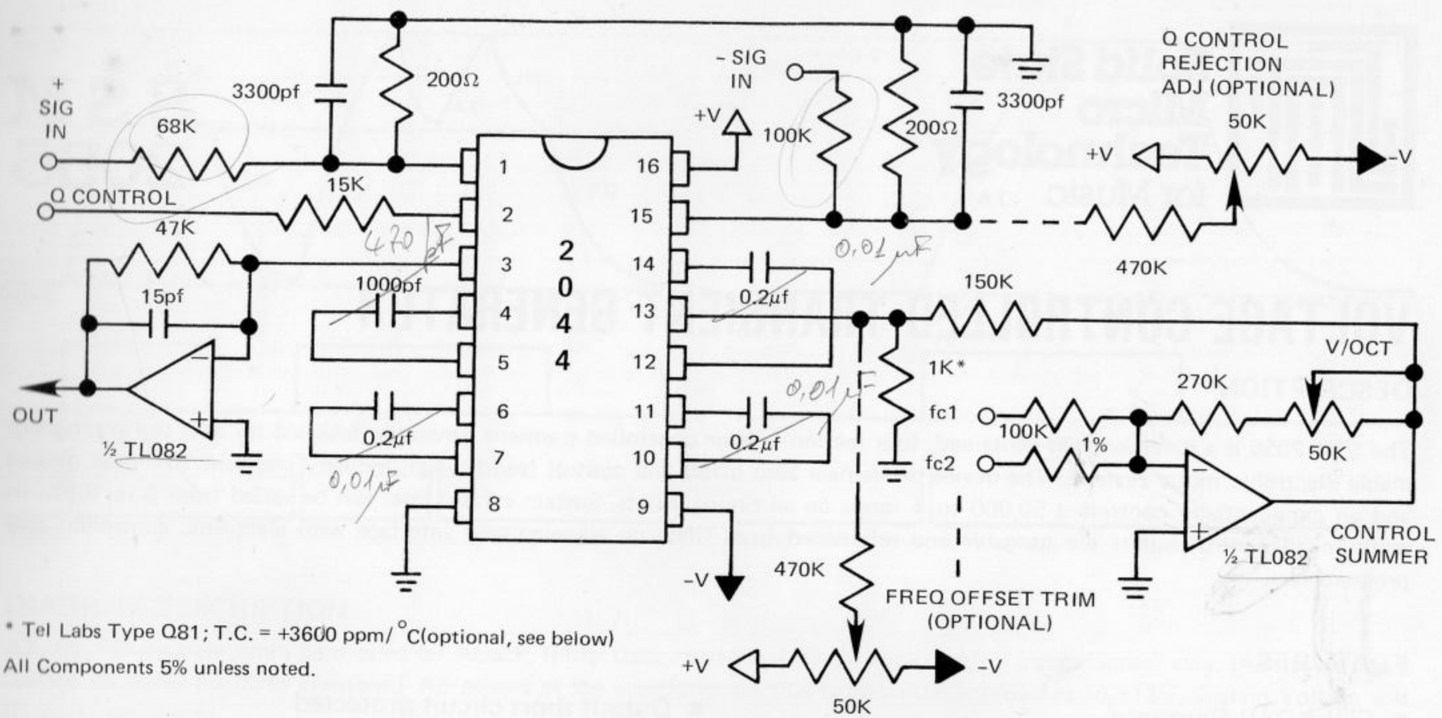


Figure 2

30VPP



* Tel Labs Type Q81; T.C. = +3600 ppm/°C(optional, see below)
All Components 5% unless noted.

Figure 3—Typical Connection

The figure above shows the typical connection of the 2044 as a four-pole lowpass electronic music filter. The differential signal inputs will accept any signal(s) up to ±18V peak-to-peak. If two oscillators are used in a voice, the output of the second should go to the opposite filter input from the first with a 3db signal level difference. This can be accomplished by scaling the input attenuators as shown, thus preventing cancellation as the oscillators phase with each other. The output op amp is required to convert the output current to a buffered voltage. The capacitors at the input pins provide stable resonance over sweep frequency.

The sense of the Q control is from GND up with minimum resonance at GND. Oscillation will occur when the current into the Q pin reaches approximately 425 µA. With the input resistor shown this corresponds to +7.5V. The Q control rejection trim is optional and is used to minimize Q control feedthrough which is about -60db when trimmed.

The control summer adds the voltages from various control sources such as the f_c panel control, transient generator, LFO etc. Any number of signals can be summed by applying them through resistors to the summing node of the op amp. The frequency offset adjust is required in polyphonic and programmable systems to make the filter(s) sound the same for an identical input control voltage. For best control rejection, the control summer and input attenuator should be designed so that the maximum swing at the 2044 control pin corresponds to the extremes of the intended sweep range when the control summer is driven to the supplies. With the values shown, one will obtain ±90mV at the input pin which corresponds to a 1000-to-1 sweep range for ±15V supplies.

The V/octave trim and the Tel Labs temperature compensating resistor are required in applications where the filter has to produce accurate musical intervals when in oscillation. If this is not necessary the control op amp feedback network and the Tel Labs resistor can be replaced by 1% 300K and 1 K resistors respectively.

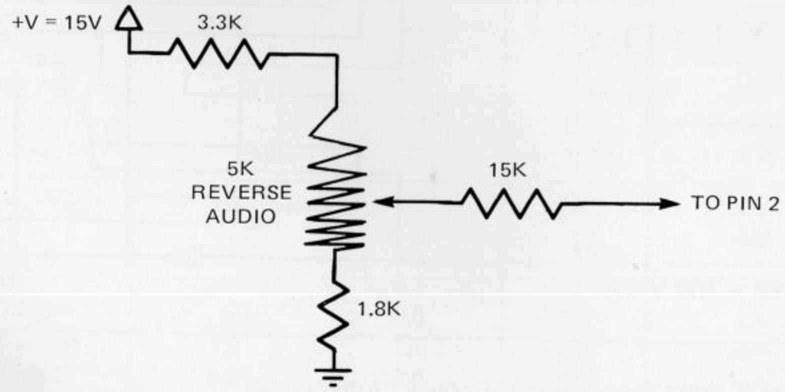


Figure 4

Solid State Micro Technology for Music

SSM 2055

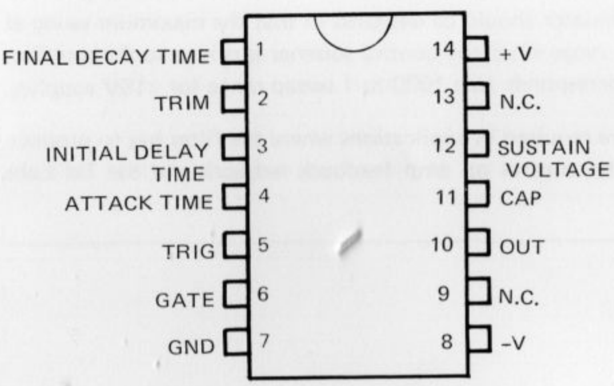
VOLTAGE CONTROLLED TRANSIENT GENERATOR

DESCRIPTION

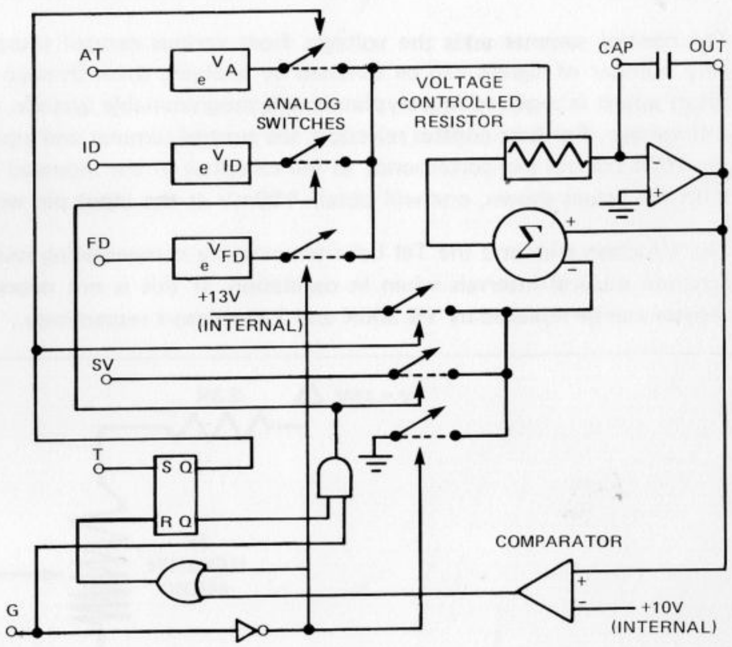
The SSM 2055 is a precision, self-contained, four section voltage controlled transient generator designed for easy use in programmable electronic music systems. The device offers near zero offset and control feedthrough, industry standard 10V peak output and an exponentially controlled 50,000 to 1 range on all timing inputs. Sustain voltage level can be varied from 0 to 100%. In addition all control inputs are gangable and referenced from GND up allowing easy interface with electronic controllers and programmers.

FEATURES

- Full ADSR Response
- Low Cost
- Independent Gate and Trigger
- Minimum external component count
- True RC contour
- ± 15 V supplies
- Small fixed final decay and sustain voltage offset
- Output short circuit protected
- Negligible control feedthrough
- Industry standard 10 V peak output
- Gangable control inputs
- Minimum 50,000 to 1 exponential time control range
- All input controls positive going from GND
- Output can drive heavy RC loads without degrading performance



PIN DIAGRAM – TOP VIEW



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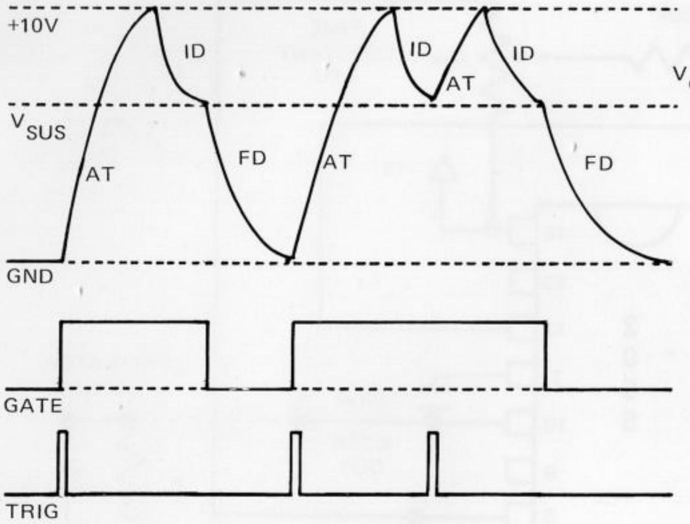


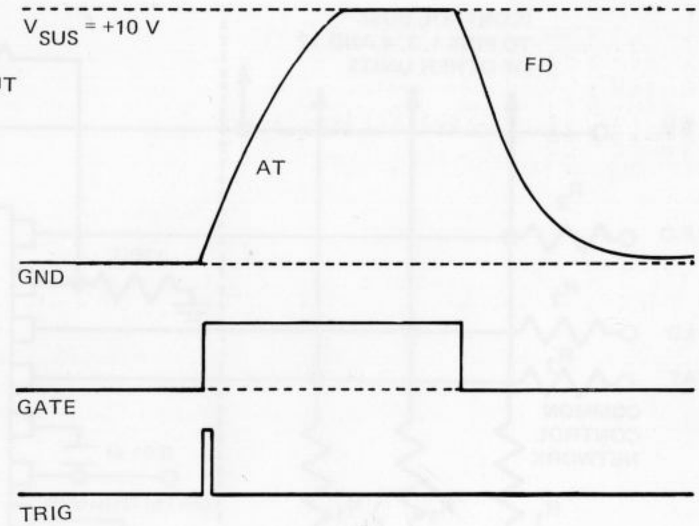


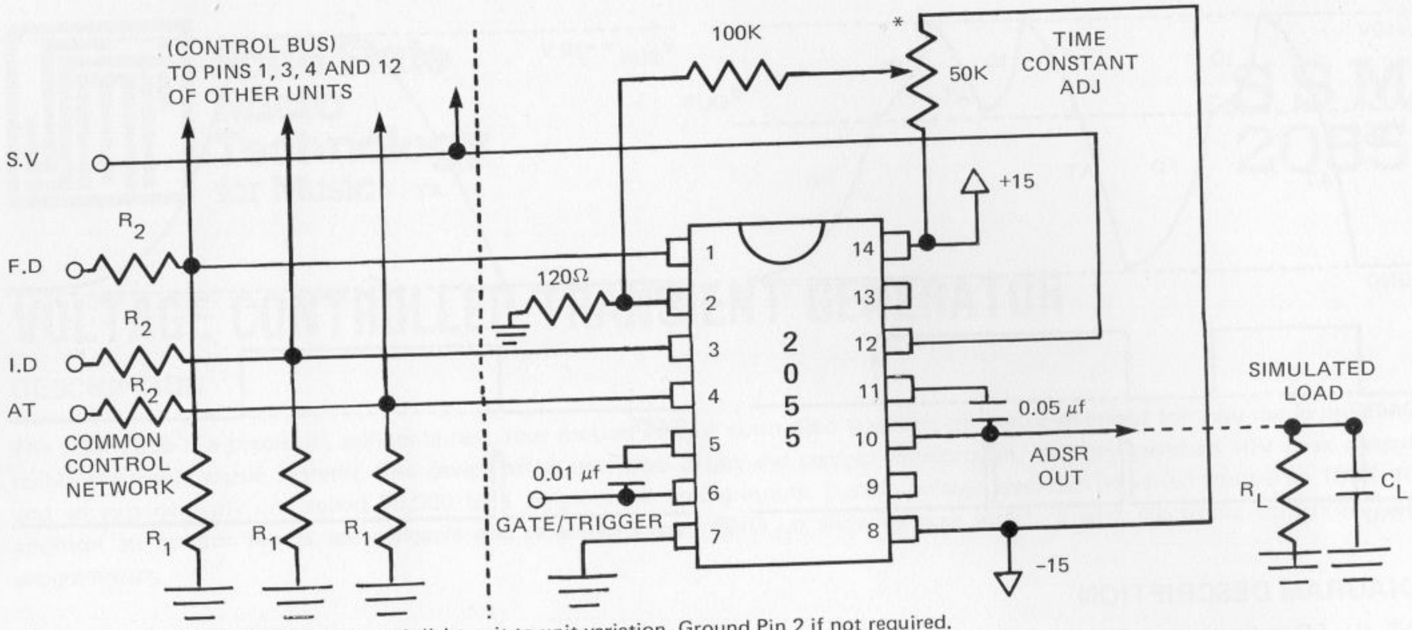
DIAGRAM DESCRIPTION

AT, ID, FD indicate times controlled by Attack, Initial Decay and Final Decay Time Control inputs respectively. (A positive going voltage increases the time constant.) All phases of the waveform are true exponential approaches to +13V, Sustain Voltage and Ground respectively.

ELECTRICAL SPECIFICATIONS

@ $V_S = \pm 15$ and $T_A = 25°C$

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
Positive Supply Current	4.00	5.8	9.00	mA	
Negative Supply Current	6.0	8.65	12.0	mA	
Positive Supply Range	+5	+15	+18	V	
Negative Supply Range	-4	—	-18	V	
Gate Threshold Voltage	1.0	1.15	1.3	V	
Gate Current (High)	-1	0	+1	μA	$V_G > 1$ V
Gate Current (Low)	—	-14	-40	μA	V_G = GND
Trigger Threshold Voltage	1.0	1.65	2.0	V	
Trigger Input Inpendance	5.14	6.85	8.56	KΩ	$V_T > 0.8$ V
Trigger Threshold Current	50	130	175	μA	
Time Control Range	50,000:1	250,000:1	—		
Time Control Sensitivity	—	+ 60	—	mV/Decade	
T.C. of Time Control Sensitivity		+ 3300		ppm/C°	
Unit to Unit Time Constant Variation (untrimmed)	0.75	1	1.30		
Attack Voltage Asymtote	12.7	13	13.3	V	
Peak Attack Voltage	9.50	9.75	10.0	V	
Attack C. V Feedthrough	—	0.5	2.5	mV	$0 < V_{AT} < 240$mV
I.D. C. V Feedthrough	—	0.5	2.5	mV	$0 < V_{ID} < 240$mV
F.D. C. V Feedthrough	—	3	13	mV	$0 < V_{FD} < 240$ mV
Sustain Voltage Offset V_o-V_{SV}	-3	-8	-13	mV	At Final Value
Final Decay Offset V_o	-1	-8	-16	mV	At Final Value
Integrator Input Current	—	0.1	1	n A	
Sustain Voltage Input Current	—	0.5	1.4	μA	
Available Output Sink Current	1.2	1.6	2.0	mA	
Positive Output Short Circuit Current	4.0	6.5	10.0	mA	



*Optional: Adjusts for capacitor value and slight unit to unit variation. Ground Pin 2 if not required.

TYPICAL CONNECTION

The diagram above shows the typical connection for a polyphonic system. The control attenuators on the left are common to all 2055's used for the same function within a voice; such as control of the final VCA. The sense of the control is from Ground up with minimum time periods at GND and increasing times at positive voltages. Some recommended resistor values for often-used sensitivites are given along with the general design equations below. The temperature coefficient of the time sensitivities can be compensated by using Tel Labs type Q81C resistors for the R_1's.

The time constant adjustment is necessary in polyphonic systems to make all voices sound the same for long attack times. The procedure is to set the AT control for the longest required attack time, ground I.D., F.D. and S.V., and adjust each 2055 to give exactly the same attack period; 10 to 20 seconds is about the longest that is musically useful. The adjustment can be ignored in manually controlled monophonic systems.

The Gate/Trigger input(s) can be driven directly from the outputs of all TTL and CMOS logic families. The ADSR output can drive any grounded load $R_L > 2.5K$, $C_L < 5000pf$.

Design Equations

$$t_A = 0.5m\ \text{sec}^{\dagger}\left\{e^{+\frac{V_a R_1 q}{(R_1 + R_2)\,kT}} + 1\right\}$$

$$t_{ID} = 0.5m\ \text{sec}^{\dagger}\left\{e^{+\frac{V_{ID} R_1 q}{(R_1 + R_2)\,kT}} + 1\right\}$$

$$t_{F.D.} = 0.5m\ \text{sec}^{\dagger}\left\{e^{+\frac{V_{FD} R_1 q}{(R_1 + R_2)\,kT}} + 1\right\}$$

$$\frac{kT}{q} = 26\text{mV @ } 25°\text{C}$$

Design Table

Input Sensitivity	R_1	R_2
1V/Decade	60Ω††	940 Ω
1V/Octave	60Ω††	3.3kΩ
1V/Decade	100Ω	1.5kΩ
1V/Octave	100Ω	5.4kΩ
1V/Decade	250Ω	3.9kΩ
1V/Octave	250Ω	13.6kΩ

NOTES

† Nominal Time Period with $V_{AT} = V_{ID} = V_{FD} = 0V$ and $C_T = 0.05\mu f$ is 1m sec.

†† Tel Labs Type Q81C = 60 Ω @ 25°C

R_1 should be kept as small as possible when the control attenuator is driving many units.

SECTION IV
SSM REPRINTS

Solid State Micro Technology for Music

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ADDITIONAL DESIGN IDEAS FOR VOLTAGE-CONTROLLED FILTERS*

by Bernie Hutchins

You have perhaps noticed that filters are popular – people love to know about them and use them. I have often wondered just why this is that filters seem to have a general appeal, and I am not sure, even though I share this love of filters. It perhaps has something to do with our feeling that all filters (not just electrical) are for the purpose of blocking out something unwanted, and somehow making the world a better place! We learn this when we sift sand for the first time as children. Here we will be taking a look at some more filter ideas, knowing that this will be a popular article, as all filter articles are. I wonder though if filters are always the most efficient way of getting the results we want in electronic music. It seems that when they are used as timbre modulators, there are simpler ways of doing things. We perhaps should look at timbre modulation from a more general viewpoint in a later issue.

Here, we want to tie up a few loose ends regarding four-pole filters, and to look at ways of applying the new Solid State Music SSM2040 IC. We will also be looking at VCF's from the point of view of the front panel (panel space) and from a human engineering viewpoint.

CORNER PEAKING OF A FOUR POLE FILTER

The four-pole low-pass filter is an old friend by now. It is often used in modular synthesizers, and may be the only filter used in some prepatched performance oriented synthesizers. We know about all we need to know about how the basic filter works. See for example the discussion in EN#41, July 10, 1974 (reprinted in the *Musical Engineer's Handbook*, Chapter 5d). The conventional form of the four-pole filter consists of four single-order sections connected in cascade. This means that the total response is fourth-order and the final roll-off rate is 24db/octave. However, the fact that the sections are individually first-order means that the corner is not going to be very sharp. This is why a regenerative path is almost always added to this filter to add resonant "corner peaking". If we were completely free with our design, we would design a sharp fourth-order filter directly, but since we want to use voltage-control, we have to be restricted to sections that are easy to control. Thus, the four cascaded first-order sections with regeneration are a very reasonable engineering compromise. Here, we will want to look at corner peaking quantitatively. Before, we have taken only a qualitative look, and this structure is too important to leave as a loose end.

The fourth-order filter is shown in Figure 1. Each one of the sections having the same transfer function, is the simple RC low-pass shown in Figure 2, and a typical voltage-controlled realization is shown in Figure 3.

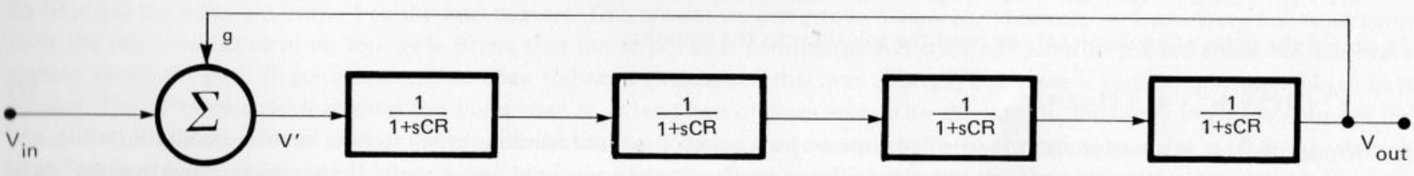

Figure 1

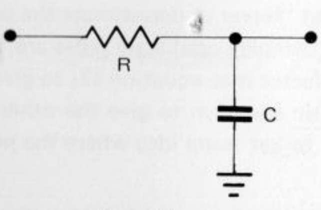

Figure 2

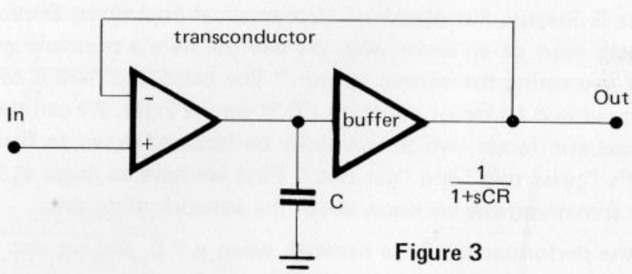

Figure 3

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With g in Figure 1 equal to zero, we have the easy case of four cascaded first-order sections, and the overall transfer function is just the product of the four sections. Thus we have:

$$T_o(s) = V_{out}/V_{in} = \frac{1}{(1 + sCR)^4} \tag{1}$$

This transfer function has a denominator that equals zero when s = –1/RC, and since there is a power of four in the denominator corresponding to this root, it is a fourth-order pole. That is, it is a fourth-order transfer function, and since there is only one pole position, there must be four poles all on top of each other at the position s = –1/RC. The pole-zero plot in the s-plane is shown in Figure 4. Note that the poles are all real, since they lie on the σ-axis. We will want to take a look at the effect of the regeneration and see how the poles move as g is made to increase in magnitude. We can say that we expect that since we get a sharper corner, some of the poles are going to have to move out into the imaginary region of the s-plane, and up closer to the jω-axis.

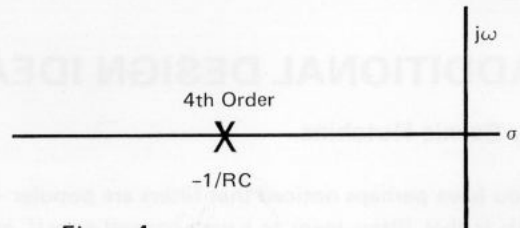


Figure 4

When g is no longer zero, we must allow for the feedback loop. First note that the original chain of four first-order sections is independent of this feedback path – it will just be processing a different input – and it is still true that:

$$V_{out}/V' = \frac{1}{(1 + sCR)^4} \tag{2}$$

Also, from the simple summing condition:

$$V' = V_{in} + gV_{out} \tag{3}$$

Equations (2) and (3) are easily combined to obtain the transfer function with feedback:

$$T_g(s) = \frac{1}{(1 + sCR)^4 - g} \tag{4}$$

At first sight, equation (4) seems to be a simple alteration of equation (1), but we shall see that things are actually very much different. While equation 1 gives a denominator in a nicely factored and useful form, equation (4) is not factored and we have at the moment no way of knowing where its poles are. First, we have to "unfactor" (multiply out) the power of four in the denominator of equation 4 so that we can include the –g term inside. It will also save a lot of notation to measure frequencies in terms of units of 1/RC, so we can set RC = 1. With this, the denominator of equation (1) becomes:

$$(1 + s)^4 = s^4 + 4s^3 + 6s^2 + 4s + 1 \tag{5}$$

and the denominator of equation (4) becomes:

$$(1 + s)^4 - g = s^4 + 4s^3 + 6s^2 + 4s + (1-g) \tag{6}$$

To obtain the poles of equation (4), we need the solutions to the equation

$$s^4 + 4s^3 + 6s^2 + 4s + (1-g) = 0 \tag{7}$$

Unfortunately, this is a non-trivial "quartic" (4th power) equation. Standard handbooks tell us how to solve quadratic, cubic, and even quartic equations, but the quartic solution is fairly involved, and you could easily spend hours doing "bookkeeping" type algebra only to end up with useless results due to a mistake somewhere up the line. [Look up the procedure in a reference such as Abramowitz & Stegun, *Handbook of Mathematical Functions,* Section 3.8.3 if you want to see how much of a mess you might be in for.] There must be an easier way. We will use here a principle we have used before called "Never underestimate the power of knowing or suspecting the correct answer." The basic idea here is to guess what two of the complex conjugate poles are, and this gives us a quadratic factor of equation (7) if we are right. We can then divide this quadratic factor into equation (7) to give the remaining quadratic factor, which can easily be factored down to first order with the quadratic equation to give the other pair of poles. That's "guess two" and "get two". First we have to guess at least one set of answers to get some idea where the poles are. We will use something else we know about the network to do this.

We know the performance of the network when g = 0, and we also know the general effects of corner peaking. Happily, we also know something about the extreme case of corner peaking – the network will oscillate. In order to see how and why the network can be made to oscillate, we need to consider briefly in review the response of a first-order section. That is, we want to consider the frequency response and phase response of a section with transfer function:

$$T(s) = 1/(1 + sCR) \tag{8}$$

To get the frequency response, we substitute $j\omega$ for s in equation 8, and take the magnitude of T(s), $|T(s)| = [T(j\omega)\cdot T(-j\omega)]^{1/2}$. This gives:

$$|T(s)| = \left[\frac{1}{1+j\omega RC}\cdot\frac{1}{1-j\omega RC}\right]^{1/2} = \left[\frac{1}{1+\omega^2R^2C^2}\right]^{1/2} \tag{9}$$

It is clear that when $\omega = 1/RC$, $|T(s)| = 1/\sqrt{2}$. Also, the phase response is given as the inverse tangent of $\omega/(1/RC)$ which will amount to 45° at $\omega = 1/RC$. Now, it is only necessary to make again a point just made above – the string of four first-order sections is independent of the loop – it always acts as a simple four stage low-pass filter. Thus, for four stages at frequency 1/RC, the total loss is $(1/\sqrt{2})^4 = 1/4$, and the total phase shift is $4\cdot 45° = 180°$. Hence, a gain of 4 and a 180° inversion is all we will need to sustain oscillation.

This is the one extra data point we need to get going. We know that when the gain g = –4, we get oscillation at 1/RC. Since we are measuring frequencies in terms of 1/RC in the main part of this analysis, this means that the oscillation frequency is 1. An oscillation corresponds to two poles that set on the $j\omega$ - axis at the frequency of oscillation and its negative value. Thus we expect that a gain of –4 results in a pair of poles at +j and –j. Substituting g = –4 into equation 7 gives:

$$s^4 + 4s^3 + 6s^2 + 4s + 5 = 0 \tag{10}$$

Since we say there should be poles at +j and –j, then (s – j) and (s + j) should be factors of equation (10), and the product (s–j) (s+j) $= s^2 + 1$ also is a factor. Thus, we should be able to divide equation (10) by $s^2 + 1$ and come out even*. In fact, if we do so, the result is $s^2 + 4s + 5$. This in turn can be factor by the quadratic formula into (s + 2 – j) (s + 2 + j). In summation, we can rewrite equation (10) as:

$$(s - j)\ (s + j)\ (s + 2 - j)\ (s + 2 + j) = 0 \tag{11}$$

We can now list all four poles of the oscillating network:

s = +j
s = –j
s = –2 +j
s = –2 –j

We can now plot the poles for two extreme cases of feedback: no feedback (g = 0) and feedback for g = –4, corresponding to oscillation. These pole positions are shown in Figure 5. We note that as expected, there are poles on the $j\omega$ axis that cause oscillation. The new thing we learn is the position of two other poles in the case of maximum feedback. Clearly all four poles have separated from the –1 point and split to four different corners in a symmetric pattern. We have not yet determined how the poles migrated from –1 to their new positions. Again, just show the migration by wavy lines as a confession of ignorance.

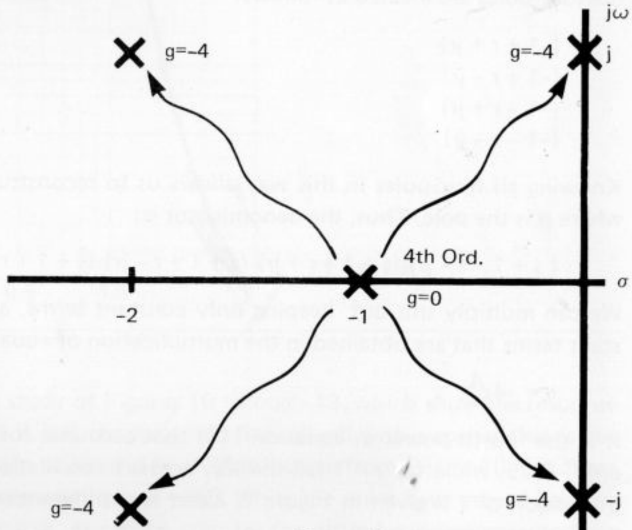


Figure 5

So how did the poles get from –1 to the four corners? One answer we can give is "smoothly" because we know from our experience with the resonance control on four-pole filters that the response is continuous and does not jump in sudden steps. We can make guesses about the path (I guessed first that they followed circles, but that was wrong.), but there is another way available to us if needed. This is to observe that since the poles start at –1, and two of them end up on the $j\omega$-axis, then they must have crossed the line $\sigma = -0.5$ (or any similar choice). Thus, there must be a solution (for a certain value of g) that places poles at $-0.5 \pm bj$, where b is not known yet. Such a pair of poles result in a quadratic factor $s^2 + s + (0.25 + b^2)$. We can then divide this quadratic factor into equation (7). When we get to the bottom of the division, terms in s must cancel as must the constant term, and since we can adjust b as needed, and g is also a free choice, this can be done. This is illustrated below:

$$\begin{array}{r l}
 & s^2 + 3s + (2.75 - b^2) \\
s^2 + s + (0.25+b^2) \;\big) & s^4 + 4s^3 + 6s^2 + 4s + (1 - g) \\
 & \underline{-s^4 - s^3 - (0.25+b^2)s^2} \\
 & 3s^3 + (5.75-b^2)s^2 + 4s \\
 & \underline{-3s^3 - 3s^2 - 3(0.25+b^2)s} \\
 & (2.75-b^2)s^2 + (3.25 - 3b^2)s + (1-g) \\
 & \underline{-(2.75-b^2)s^2 - (2.75 - b^2)s - (2.72-b^2)\ (0.25+b^2)} \\
 & (0) \qquad (?) \qquad (??)
\end{array}$$

*If you have forgotten how to do this, see AN-57.

In order for the s term to cancel out, it is necessary that:

$$3.25-3\,b^2-2.75+b^2=0.50-2b^2=0 \quad \text{or } b=0.5 \tag{12}$$

In order that the constant term cancel out, it is necessary that:

$$(1-g)-(2.75-b^2)\,(0.25+b^2)=(1-g)-(2.5)\,(0.5) \quad \text{or } g=-0.25 \tag{13}$$

With b determined as 0.5, we place our first set of poles at $-0.5 \pm 0.5j$. Also, the second quadratic factor just determined above is $s^2 + 3s + 2.5$, which by the quadratic formula gives poles at $-1.5 \pm 0.5j$. These poles lie on the straight lines that would connect the original poles at -1 to the final poles in the corners in Figure 5. That is, it appears that the wavy lines we drew should really have been straight. By repeating the procedure, it is possible to show that the lines (pole loci) are really straight (or that we are very unlucky!).

The migration of the poles is now determined at least in that we know the paths and the limits of these paths. What remains to be done is to find a relationship between pole position and the feedback factor g at points other than those we have just tested individually. Eventually, from the pole positions we expect to learn about the characteristics of the frequency response.

To determine the exact relationship between the feedback factor g and the pole positions, we first observe that we have shown that the poles move out from -1 in a square pattern. It is convenient to define a new "r" (resonance) variable to measure the size of the square as in Figure 6. The value of 4 runs from 0 to +1 as go goes from 0 to -4. In terms of r, the four poles are located as follows:

$(-1 + r + jr)$
$(-1 + r - jr)$
$(-1 - r + jr)$
$(-1 - r - jr)$

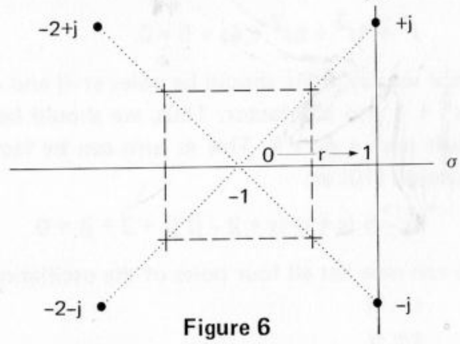


Figure 6

Knowing all four poles in this way allows us to reconstruct the denominator of the transfer function as the product of all (s – p) where p is the pole. Thus, the denominator is:

$$(s+1-r-jr)\,(s+1-r+jr)\,(s+1+r-jr)\,(s+1+r+jr) \tag{14}$$

We can multiply this out, keeping only constant terms, and comparing with equation (6), this can be set equal to $(1-g)$. The constant terms that are obtained in the multiplication of equation (14) are $4r^4 + 1$. Hence $1-g = 1 + 4R^4$ and:

$$g=-4r^4 \tag{15}$$

It is the fourth power in equation (15) that accounts for the fact that in our example with poles at -0.5 (half way across), g was only -0.25, while for $r = 1$ (all the way across to oscillation) g had to reach -4. For convenience, a plot of the required value of g as a function of r is given in Figure 7. Later we will want to look at this when selecting a means of controlling the value of the feedback g in a practical module.

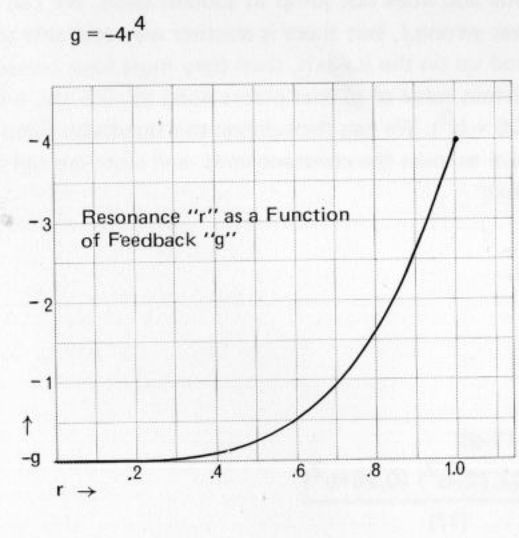


Figure 7

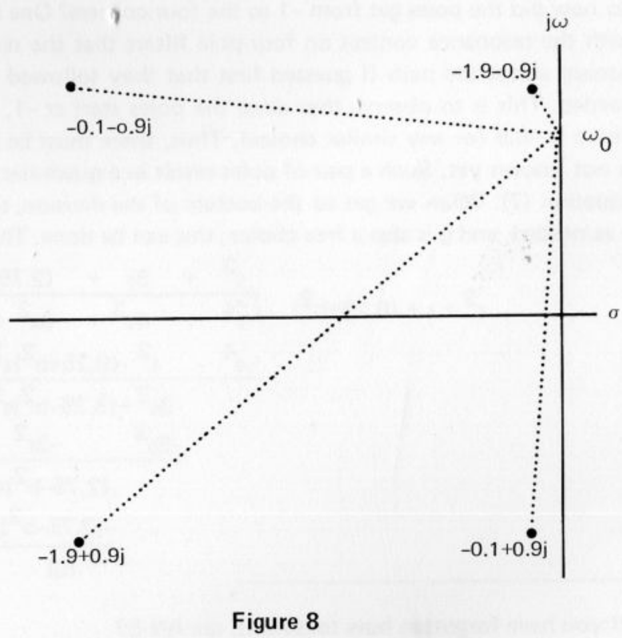


Figure 8

The frequency response of the four-pole filter with feedback can be determined from the pole positions which we have just determined. For example, with g = -2.63, the poles are positioned at r = 0.9 as shown in Figure 8. A graphical method can then be used to determine the frequency response. First we select any one point where we want to know the magnitude of the frequency response (for example, ω_0 in Figure 8. We then measure the distances from the poles to the point ω_0. These are shown as dotted lines in Figure 8. The final step is to divide 1.00 by the product of these four distances. This is repeated until enough points are obtained to sketch in the full curve. For more information on this method, see AN-45.

Figure 9 shows a comparison of a theoretical (graphically calculated) frequency response and an experimentally measured response corresponding to the pole positions shown in Figure 8. Note that the high peak at a frequency of 0.9 is due mainly to the close approach of the pole at -0.1+0.9j to the $j\omega$ axis. The peak shown illustrates the meaning of "corner peaking." Note that the peaking is quite extensive (this is a log plot) and in many ways the response resembles a bandpass response.

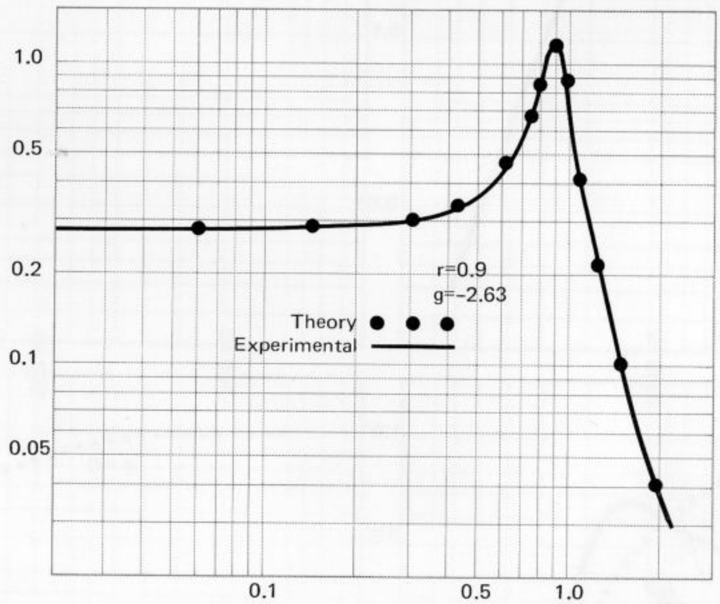


Figure 9

A better feeling for the effect of corner peaking can be obtained from a study of Figures 10 through 13, which show the response for increasing values of r. Figure 10 shows the response when g = 0, and is just the expected flat roll-off with a corner that is not too impressive compared to a Butterworth corner, for example. We can begin to see the final 24db/octave from Figure 10, but there is still a ways to go. Figure 11 shows the effect of feedback of -0.25 corresponding to r = 0.5. While the poles have moved half the way to their oscillation positions, there is relatively little change in the curve, as can be seen by the dotted curve represented the original g = 0 curve. Note that feedback does lower the DC response level. Figure 12 shows the curve for a value of r = 0.75 (g = -1.26) and we see a much improved corner with some ripple and overall gain loss. Probably, this reminds us of a second-order Chebyshev as much as anything, and in fact, with two poles moving to the rear, the close-in part of the curve is very similar to a two-pole system. Figure 13 shows even more feedback (g = -2.63, r = 0.9) and a response that is looking less and less low-pass and more and more bandpass. Beyond r = 0.9, we are mainly working with a high-Q response very much like a bandpass.

From Figures 10-13, we can see that there is relatively little change that we might expect to become important musically until a value of r that approaches 0.75. This requires feedback of -1.26 or greater. The significance of this is that in some feedback schemes we are concerned with relatively small amounts of feedback, and therefore employ such things as log pots and exponential responses to obtain very fine control. Here, we are mainly interested in a range from 25% to 100% of the maximum value. A log pot, which gives 15% resistance change for the first 50% rotation would be a waste here. In fact, even the bottom 25% of a linear pot would be wasted. Thus we might even suggest a resistor in series with the pot to raise the whole thing up some, and a "reverse audio" pot would seem ideal for controlling resonances in the range of 0.9 to 1.0 where things change very rapidly as far as the response is concerned. Some suggested feedback circuits thus appear as seen in Figure 14. In the first, a resistor has been added to a linear pot. In the second, a bypass resistor is used (assuming feedback to a summing node). The final circuit uses a reverse audio pot to control feedback. The reverse audio pot is a backward log pot – most of the resistance change is in the first 50% of clockwise rotation.

SUMMARY OF FINDINGS: The effect of the feedback or resonance control of the usual form of the traditional electronic music four-pole filter is to spread the four poles, which are initially piled up at -1/RC, out in a square pattern. The frequency response of the filter is understood in terms of the new pole positions. While a small amount of feedback results in a substantial displacement of the poles (Figure 7), this large displacement has a relatively small effect on the frequency response curves (Figure 11). Large amounts of feedback result in relatively little additional pole displacement, but more substantial changes in the frequency response

curves (Figures 12 and 13). The implication is that controls used to set the resonance of the filter should have special features to increase resolution at feedback levels between 25% and 100% maximum (just the opposite of a log pot, for example).

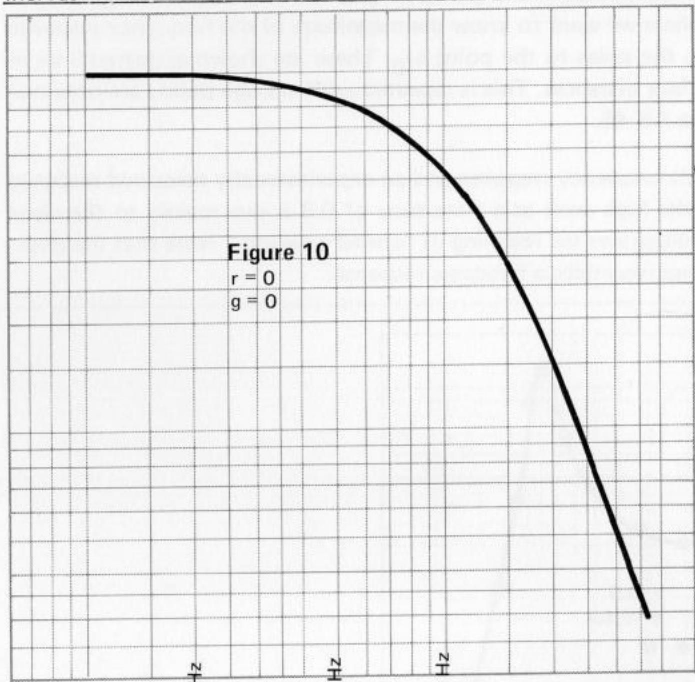


Figure 10
r = 0
g = 0

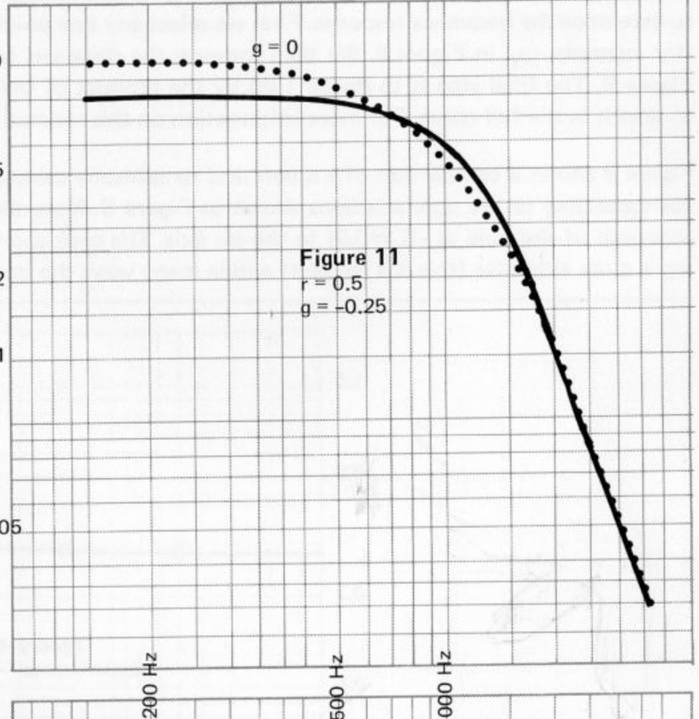


Figure 11
r = 0.5
g = -0.25

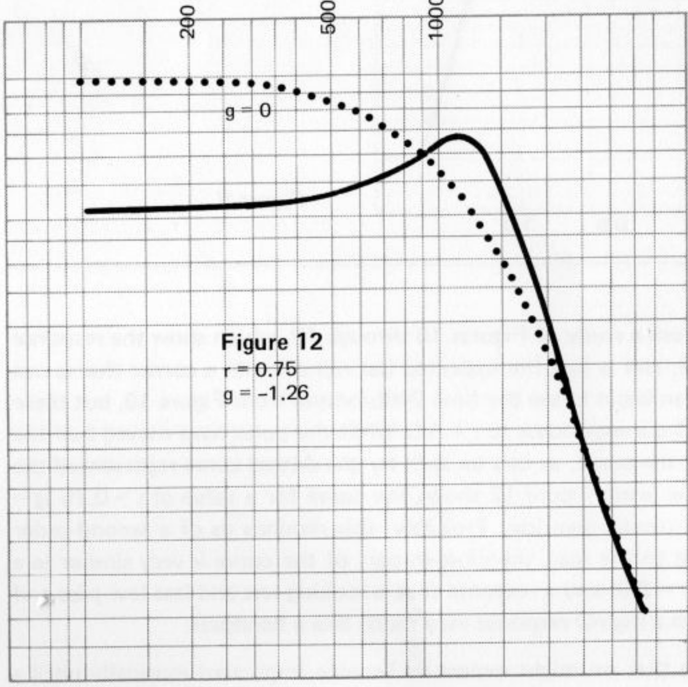


Figure 12
r = 0.75
g = -1.26

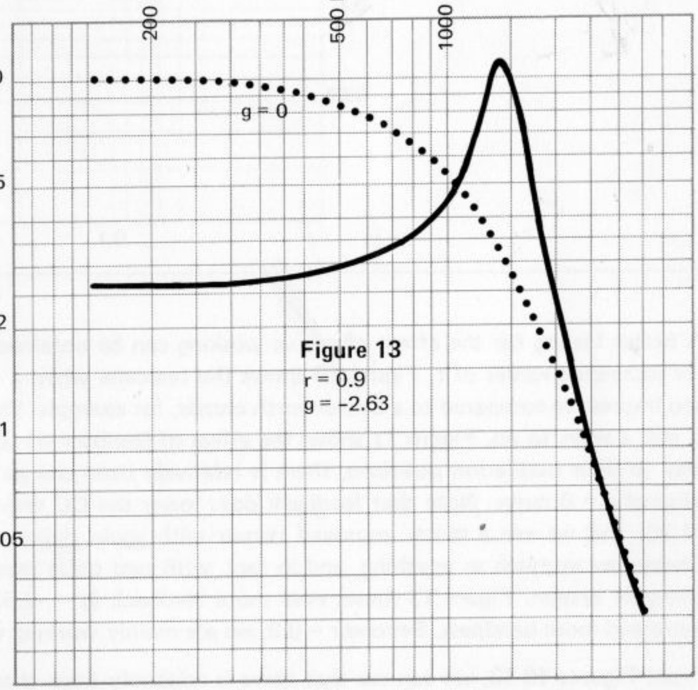


Figure 13
r = 0.9
g = -2.63

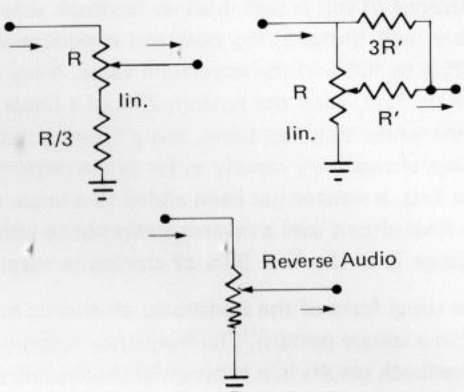


Figure 14

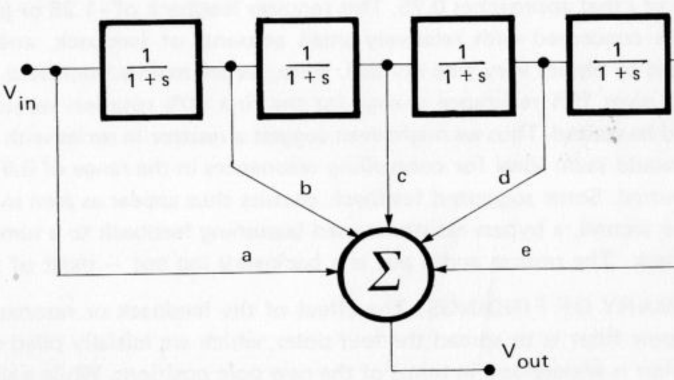


Figure 15

OBTAINING BANDPASS AND HIGH-PASS FUNCTIONS FROM A FOUR-POLE LOW-PASS:

One of the big "selling points" of the state-variable VCF is that it provides several modes of operation (low-pass, bandpass, high-pass, and notch) as simultaneous outputs. Here we will show that it is possible and easy to obtain a full range of functions from the basic four-pole low-pass. These functions are fourth-order in this case, although the method is general and can give third, second, or first-order as well.

It is obvious that we can "tap" any of the four first-order sections of the four-pole low-pass to obtain a first, second, third, or fourth-order low-pass response. What we will show is that it is possible to do a weighted sum of all four taps plus the input to obtain a general fourth-order numerator for the transfer function, retaining the original fourth-order denominator. The summing network is indicated in Figure 15. From the diagram, we get:

$$T(s) = V_{out}/V_{in} = a + b/(1+s) + c/(1+s)^2 + d/(1+s)^3 + e/(1+s)^4 \tag{16}$$

$$= \frac{as^4 + (4a+b)s^3 + (6a+3b+c)s^2 + (4a+3b+2c+d)s + (a+b+c+d+e)}{(1+s)^4} \tag{17}$$

It is clear that by properly selecting values for a, b, c, d, and e, we can obtain any fourth-order numerator we want. A trivial example is when a, b, c, and d are zero and e = 1, in which case, we have our fourth-order low-pass back. We will want to look at how we can obtain the necessary weightings, and then will want to examine the effect of corner peaking of the low-pass "backbone".

Probably the most interesting case we have is the conversion to a four-pole high-pass. To get a fourth-order high-pass, we need to get an s^4 in the numerator, and all else must go. You can set up equations if you wish, but you can probably just see that a should equal 1 and 4 a+b should equal zero, hence b = -4, and so on. The results for a high-pass are:

$$a = 1 \qquad b = -4 \qquad c = 6 \qquad d = -4 \qquad e = 1 \tag{18}$$

which converts equation (17) into:

$$T(s) = \frac{s^4}{(1+s)^4} \tag{19}$$

To test the theory, we can use a non-voltage-controlled setup of the type shown in Figure 16. The first-order sections have transfer function $-1/(1+s)$. It is simplest to just sum into an inverting summer, and the inversion in the transfer function automatically makes every other coefficient negative, exactly what we need.

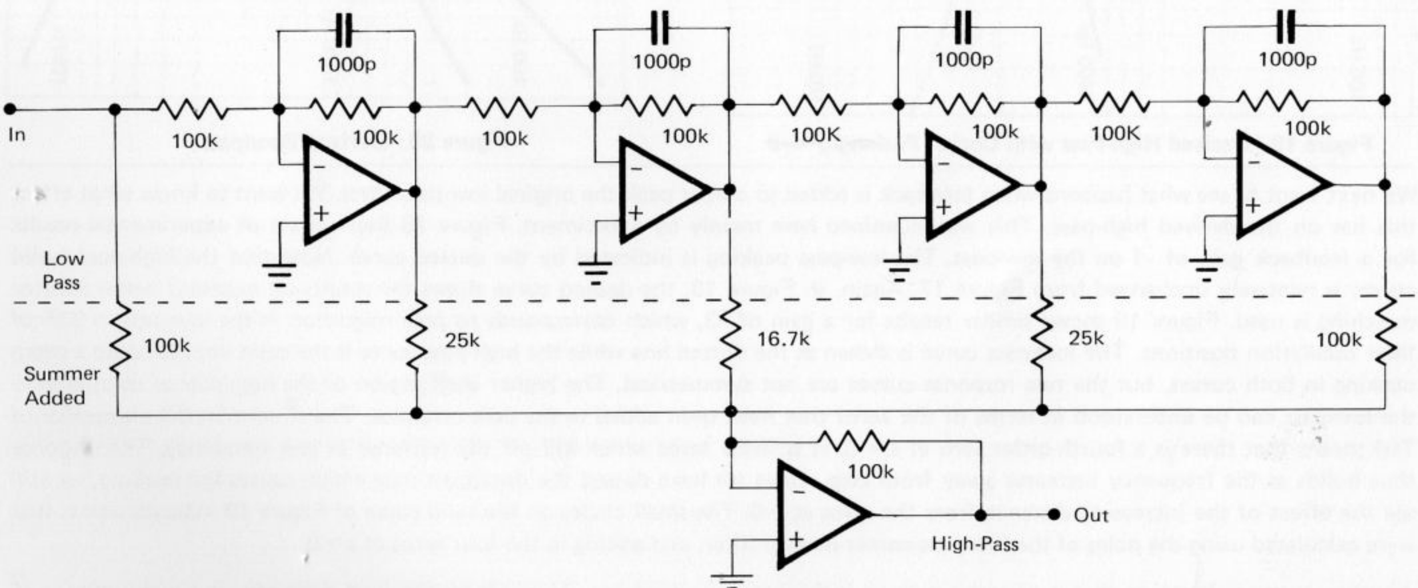


Figure 16. Test Circuit of Figure 15 for High-Pass Mode.

Figure 17 shows experimental data on the filter of Figure 16. The dotted line shows the low-pass output of the original section. The solid line is the experimental data on the high-pass output while the dashed line shows a high-pass roll-up that we would have hoped to obtain. The reason it was not obtained is almost certainly a lack of resistor precision (5% resistors were used). The solid part of the curve in itself is sufficient to confirm the basic operation of the circuit according to theory.

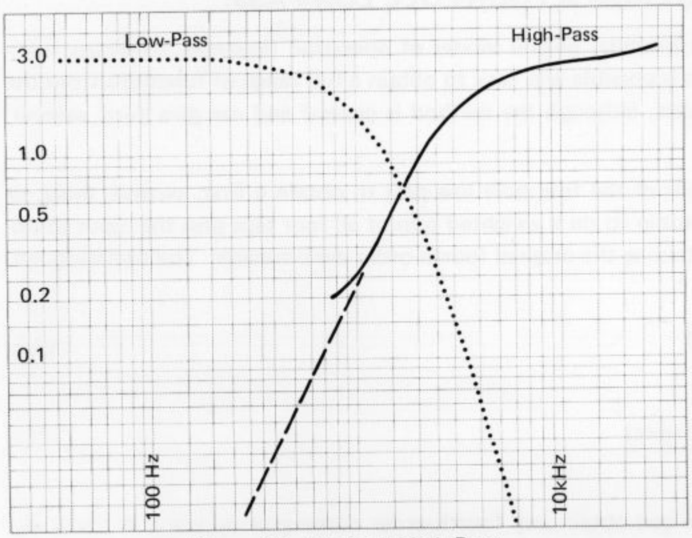


Figure 17. Derived High-Pass

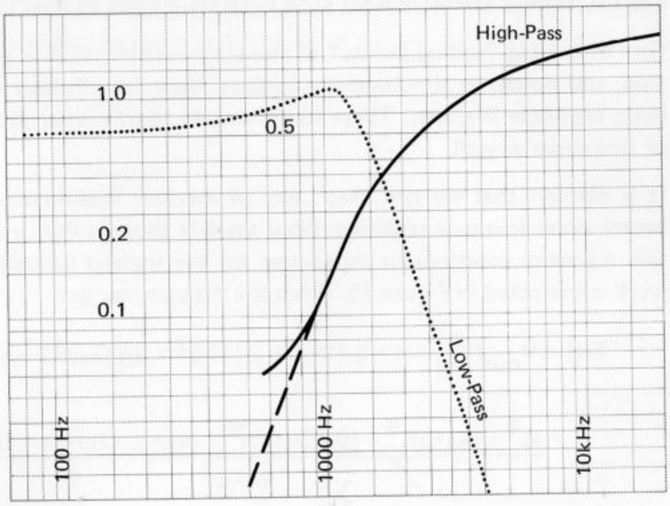


Figure 18. Derived High-Pass with Corner Peaking g = –1

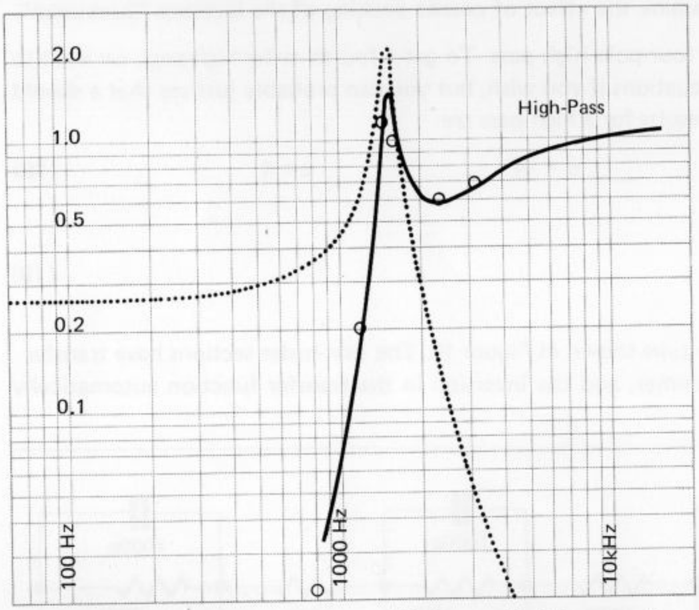


Figure 19. Derived High-Pass with Corner Peaking g = –3

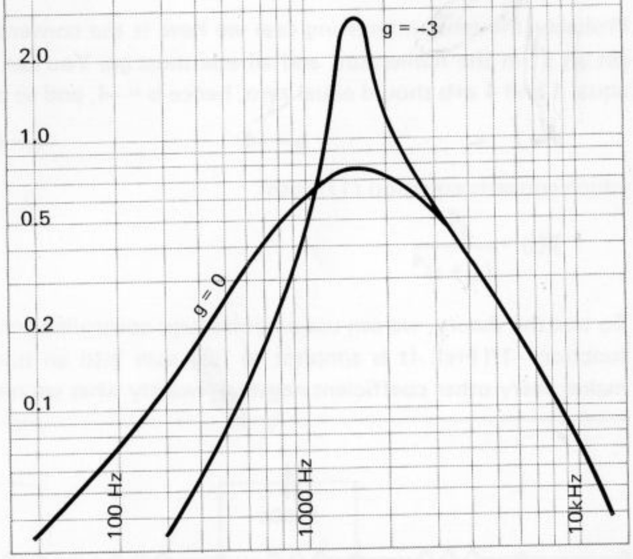


Figure 20. Derived Bandpass

We next want to see what happens when feedback is added to corner peak the original low-pass filter. We want to know what effect this has on the derived high-pass. This was examined here mainly by experiment. Figure 18 shows a set of experimental results for a feedback gain of –1 on the low-pass. The low-pass peaking is indicated by the dotted curve. Note that the high-pass (solid curve) is relatively unchanged from Figure 17. Again, in Figure 18, the dashed curve shows the results we expect if better resistor matching is used. Figure 19 shows similar results for a gain of –3, which corresponds to pole migration in the low-pass to 93% of their oscillation positions. The low-pass curve is shown as the dotted line while the high-pass curve is the solid line. We note a sharp peaking in both curves, but the two response curves are not symmetrical. The higher shelf region of the high-pass as compared to the low-pass can be understood in terms of the zeros that have been added to the pole-zero plot. The s^4 term in the numerator of T(s) means that there is a fourth-order zero at s = 0. It is these zeros which kill off the response at low frequency. The response thus builds as the frequency increases away from zero. Once we have passed the dominant pole which causes the peaking, we still see the effect of the increasing distance from the zeros at s=0. The small circles on the solid curve of Figure 19 indicate points that were calculated using the poles of the low-pass corner-peaked filter, and adding in the four zeros at s = 0.

Another response function that is of interest to us is the bandpass function. This will be obtained if we can set a numerator to s^2 in the transfer function and get rid of everything else. The set of summing parameters that will achieve this are:

a = 0
b = 0
c = 1
d = –2
e = 1

It may bother you a little that a and b are zero so only three stages are tapped, and you might wonder how the filter knows it is fourth order if it only has contact with three stages, and not five. The answer to this seems to be that there are two stages of low-pass filtering that have an effect before the taps are reached. If it were the case that the first three stages were tapped, not the last three, then the filter could not be fourth order. It is a simple matter to set up a circuit similar to that in Figure 16 to realize the necessary summing network, and experimental results are shown in Figure 20. Ignore for the moment the g = –3 curve. The shallower curve should be a bandpass rolling off each side at 12db/octave, because this is a fourth-order bandpass. Both sides are sharper than 6db/octave (a 45° angle), although it is a little difficult to be sure exactly what they are. The g = –3 curve shows the effect of corner peaking. Note that the lower side of the curve sharpens considerably while the upper side remains the same.

We have shown that the basic theory seems to work out in practice. In addition, a notch response and any number of other special responses can be obtained. A notch with second-order poles at +j and –j can be obtained with a numerator $s^4 + 2s^2 + 1$ obtained with a = 1, b = –2, c = 2, d = –2, and e = 2. However, with this notch, as with most high order notches, component precision must be very high or the notch washes out. Our experimental data with 5% resistors gave a notch only down by a factor of 4.

We should mention that this same method works for second order as well (and can probably be extended for higher orders). For second order, there would be two low-pass stages and three summing coefficients a, b, and c. For high-pass, a = 1, b = –2, and c = 1. For bandpass, a = 0, b = 1, and c = –1. For low-pass, a = 0, b = 0, and c = 1. For notch, a = 1, b = –2, and c = 2. See AN-71, to be published.

The idea of feedback to the input from the summed output rather than from the low-pass output of course comes in mind. In the high-pass case, it does not work because as we have discussed before, all high-pass filters come down somewhere, and there is usually associated phase shift which will meet all the conditions of the low-pass filter as an oscillator. Hence we get oscillation at high frequency. In the case of the bandpass, it is possible to feed back the summed (bandpass) output to get corner peaking. Our experiments showed that the results were much the same as with the low-pass peaking.

PRESENT DAY VCF DESIGN OPTIONS

For many years, the four-pole low-pass and the state-variable VCF's have been the principal choices for electronic music. Filter structures that combine the main advantages of the state-variable (multi-function) and the four-pole low-pass (fourth-order response) are attractive new options. The state-variable and the four-pole low-pass are well understood (*Musical Engineer's Handbook,* Chapter 5d, and elsewhere in back issues of this newsletter: in particular, EN#71 on the ENS-76 VCF options). We have also demonstrated that a two-section state-variable approach can be used to combine advantages of state-variable and four-pole (EN#58). In the present report, we have shown that a four-pole approach can be used to arrive at multi-functions. In many ways, we seem to have two roughly equivalent choices for developing a new type of filter. [There is also the "dual-shift" filter that was discussed in EN# 81]. In some cases, we can get help in choosing between design options by considering different amounts of hardware required, but here, all the methods we have considered require about the same amount of hardware, so we must look for more subtle differences in making a choice. It will be simplest to just tabulate the design factors that *are* different for the different approaches and then discuss some of the more important differences.

	Two State-Variable	4-Pole plus Summers	Two Dual-Shift
Principal Advantage	Low Sensitivity	Parallel Outputs	Possibly less phase shift in controlled stages
Principal Drawback	Need for Switching	High Sensitivity	Need for Switching
Response Curves	Regular Shapes	Some Irregularities	Regular Shapes
Control of Resonance	Double Control	Single Control	Double Control

First, we can look at the principal drawbacks. With the state-variable, which normally has parallel outputs (simultaneous, LP, BP, HP and notch), when we try to cascade two second-order sections, we have to switch the appropriate output of the first section to the input of the second. Thus, there is a need for a special switch (probably two-pole, 7-position, see EN #84 (17)). With the 4-pole and summers, all outputs can be parallel. The principle drawback of the 4-pole is that when summers are used and when we go to fourth-order, the component sensitivity of the high-pass, bandpass, and notch outputs is quite high, and precision components may be required. The state-variable on the other hand is quite insensitive.

A lesser drawback of the 4-pole with summers is that the response shapes are not symmetrical when resonance is added. See for example, Figure 19 of this report. A lesser drawback of the state-variable is that both sections must have individual resonance controls. Since we probably want voltage-controlled resonance with this VCF, this means another transconductor is required, or with manual control, a dual pot, which is probably more of a problem than the transconductor.

In many cases, we are very much interested in the amount of panel space that will be required for a given design. In such cases, a filter with switched modes and a single output (two positions on the panel) may be more useful than a filter with parallel outputs (up to 8 positions). This might tend to favor the two state-variable design, but of course, we can use a switch and single output with the four-pole design as well, and the switch may be simpler in the four pole case.

Another point about panel space for VCF's is that we may be able to implement our frequency controls in a manner different from what seems to be necessary for VCO designs. We may not need a fine frequency control for the VCF, although this seems essential for VCO's. Where panel space is really at a premium, a multi-turn pot may be used for frequency level, coarse and fine together.

A final point about panel space and VCF's is that the VCF is generally just going to require much more space than other modules. This is sure to be true if voltage-controlled resonance is used. Some users also find a second envelope input to a VCF useful. Thus, if you have to cut back on panel space, you must be willing to cut back on VCF features as well.

Let's take an example where you have available to build a filter four transconductors all controlled in parallel, and as many op-amps for buffers or summers as you need. This could be a SSM2040 chip and a few extra op-amps, or it could be formed from CA3080's and other individual IC's. The problem is to come up with a useful fourth-order filter. If we have no regard for panel space, there are many things we can do, and the designer probably has no problems in such a case. Here we want to look at designs that will use a small amount of panel space. Two possible setups will be described. Figure 21 shows a two state-variable approach.

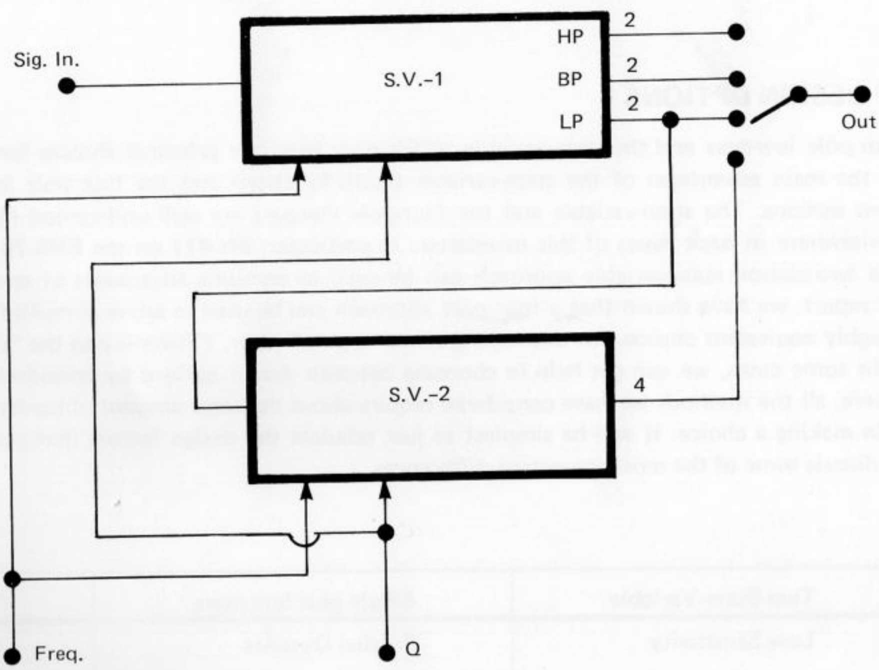


Figure 21

Here we avoid switching on the input of the second stage by just feeding in the LP from the first stage. With this arrangement, we have a regular state-variable (second-order) LP, BP, and HP, and a fourth-order LP as well. The output panel space requires only two positions, a rotary switch and a jack. The second example is seen in Figure 22 and is based on the four-pole low-pass. Here we have provided summers for second and fourth-order low-pass and high-pass. A single bandpass output (either second or fourth-order) seems sufficient because the Q of the filter is variable. A second-order notch is used because the fourth-order notch is too sensitive to component tolerance. As with Figure 21, the output panel space is only two positions, a switch and a jack.

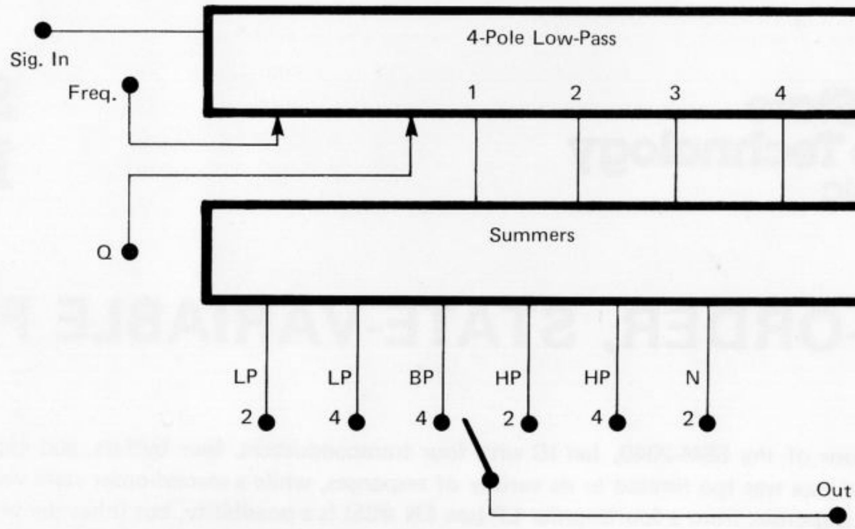


Figure 22

At some point, we have to call a halt to electrical engineering to consider some musical engineering and some human engineering. What do musicians need in filters to make their music. For musicians where timberal control is not a big part of their music, relatively little is required of the filters. Musicians who do use timber variations as an important part of their music will require a lot of their filters, as will musicians doing "imitative synthesis" of traditional sounds. But regardless of their demands, they will describe their filters in terms of a "good feel" and a "good sound" and probably only resort to terms like "four-pole" as a necessary "buzz-word" to keep the engineer's attention. Things like the shape of the frequency response curve take a back seat to the feel and the sound and the ease with which the musician is able to find the right knob. This is not an easy job by any means. Probably the most important thing that new VCF designs and VCF chips can do for us is to make the electrical engineering simple enough that we can consider some of the human engineering aspects of filter design. We can begin to do things that are not *just* reasonable electronically. In using VCF chips for example, we need not exploit all the capabilities of the device (as in Figure 21) but can use the chip because it saves effort.

Solid State Micro Technology for Music

S S M
Reprint

A FOURTH-ORDER, STATE-VARIABLE FILTER*

by Bob Chidlaw

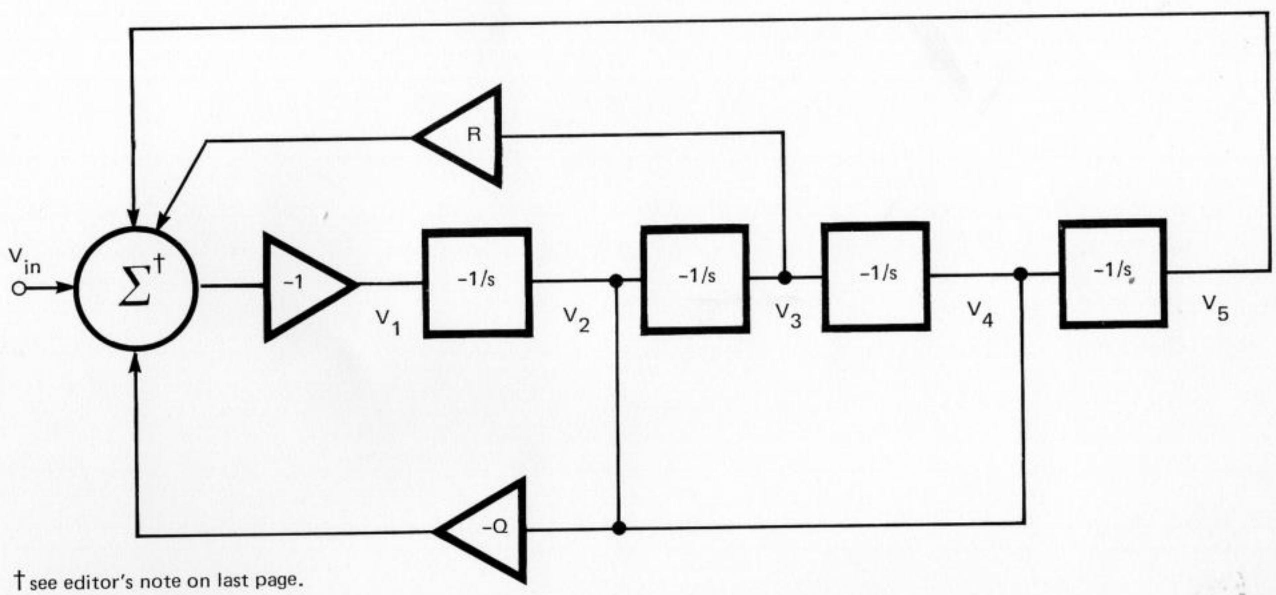
While contemplating applications of the SSM-2040, (an IC with four transconductors, four buffers, and exponential control), it seemed that a fourth-order low-pass was too limited in its variety of responses, while a second-order state variable wasted half the chip. Synthesis of HP and BP responses from a fourth-order LP (see EN #85) is a possibility, but it has the problem of component sensitivity for accurate responses, which state-variable filters avoid. The solution is clearly a fourth-order state-variable filter.

Consider the filter shown in Figure 1. It has a Q control, as in a second-order state-variable filter, and a new feedback control denoted R. These operate by changing the gains of the two feedback amplifiers. (The resonance of a second-order filter will be written "q". The q's of the fourth-order filter will vary in an inverse manner with Q.) The equation for V_1 is

$$V_1 = -V_{in} + Q[V_1(-1/s) + V_1(-1/s)^3] - R[V_1(\frac{1}{s^2})]$$

or

$$-V_1/V_{in} = \frac{s^4}{s^4 + Qs^3 + Rs^2 + Qs + 1}$$



† see editor's note on last page.

Figure 1 – Fourth-Order State-Variable Filter

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Let us define here

$$D = s^4 + Qs^3 + Rs^2 + Qs + 1$$

Factors of s are removed from the numerator as we progress down the chain of integrators, until we are left with

$$-V_5/V_{in} = 1/D$$

V_1 is some kind of high-pass response, and V_5 is a low-pass. To understand their exact behavior, the location of the zeros of D must be determined in terms of Q and R. Although D = 0 is a quartic equation, it can be solved without resorting to the general quartic formula. Take the quare of a quadratic

$$(s^2 + as + 1)^2 = s^4 + 2as^3 + (2 + a^2)s^2 + 2as + 1$$

Identifying Q = 2a, then if $R = 2 + Q^2/4$, the denominator factors into the square of a quadratic, and the LP and HP transfer functions are equal to the squares of the corresponding second-order transfer functions, as would be realized by cascading identical second-order state-variable filters, with q's of 2/Q; the HP response is

$$\left[\frac{s^2}{s^2 + (Q/2)s + 1}\right]^2$$

If $R \geqslant 2 + Q^2/4$, then the denominator factors into two quadratics which do not have identical resonant frequencies. To demonstrate this, take the product

$$[(\frac{1}{c}s)^2 + a(\frac{1}{c}s) + 1]\ [(cs)^2 + a(cs) + 1] =$$

$$s^4 + a(c + 1/c)s^3 + (a^2 + c^2 + \frac{1}{c}2)s^2 + a(c + 1/c)s + 1$$

We want

$$Q = a(c + 1/c)$$
$$R = c^2 + 1/c^2 + a^2$$

These equations yield

$$c + \frac{1}{c} = K = \sqrt{\frac{(R+2) + \sqrt{(R+2)^2 - 4Q^2}}{2}}\ ,$$

$$c = \frac{K + \sqrt{K^2 - 4}}{2}\ ,$$

$$a = Q/K\ .$$

c is a parameter indicating the amount the resonant frequencies of the two quadratic sections shift up and down from the original center frequency. For small Q (large effective q), the splitting parameter c is independent of Q. A table for the amount of splitting as a function of R is given below:

R	c	K	c^2	**Split in Octaves**
2	1.00	2.00	1.00	0
2.5	1.41	2.12	2.00	1
3	1.62	2.24	2.62	1.39
4	1.93	2.45	3.72	1.90
5	2.19	2.65	4.80	2.26
6	2.41	2.82	5.81	2.54
7	2.62	3.00	6.86	2.78
8	2.81	3.16	7.90	2.98

For large Q's, there will be less splitting. The effective q is the same for both quadratics and is K/Q. For larger splittings, q is increased. The condition $R \geqslant 2 + Q^2/4$ comes from the requirement that c is real, which implies $K \geqslant 2$. For K real we must have $R+2 \geqslant 2Q$, but this is always true if $R \geqslant 2 + Q^2/4$.

For $R < 2 + Q^2/4$, an alternative factorization is needed. Take

$$(s^2 + fs + 1)(s^2 + gs + 1) = s^4 + (f+g)s^3 + (2+fg)s^2 + (f+g)s + 1$$

Then

$$Q = f+g$$
$$R = 2+fg$$

which has solutions

$$f = \frac{Q}{2} + \sqrt{2 + Q^2/4 - R}$$

$$g = \frac{Q}{2} - \sqrt{2 + Q^2/4 - R}$$

This represents a pair of second-order filters with the same resonant frequency but different q's. Note there is an instability (oscillation) at $g = 0$, when

$$\frac{Q}{2} = \sqrt{2 + Q^2/4 - R}$$

which implies

$$R = 2.$$

This holds for any Q setting. When R approaches 2, g goes as $(R-2)/Q$, or effective $q = Q/(R-2)$. R's near 2 lead to high q's.

The motion of the pole positions as a function of Q is quite elegant, and is shown in Figure 2 for R = 3.

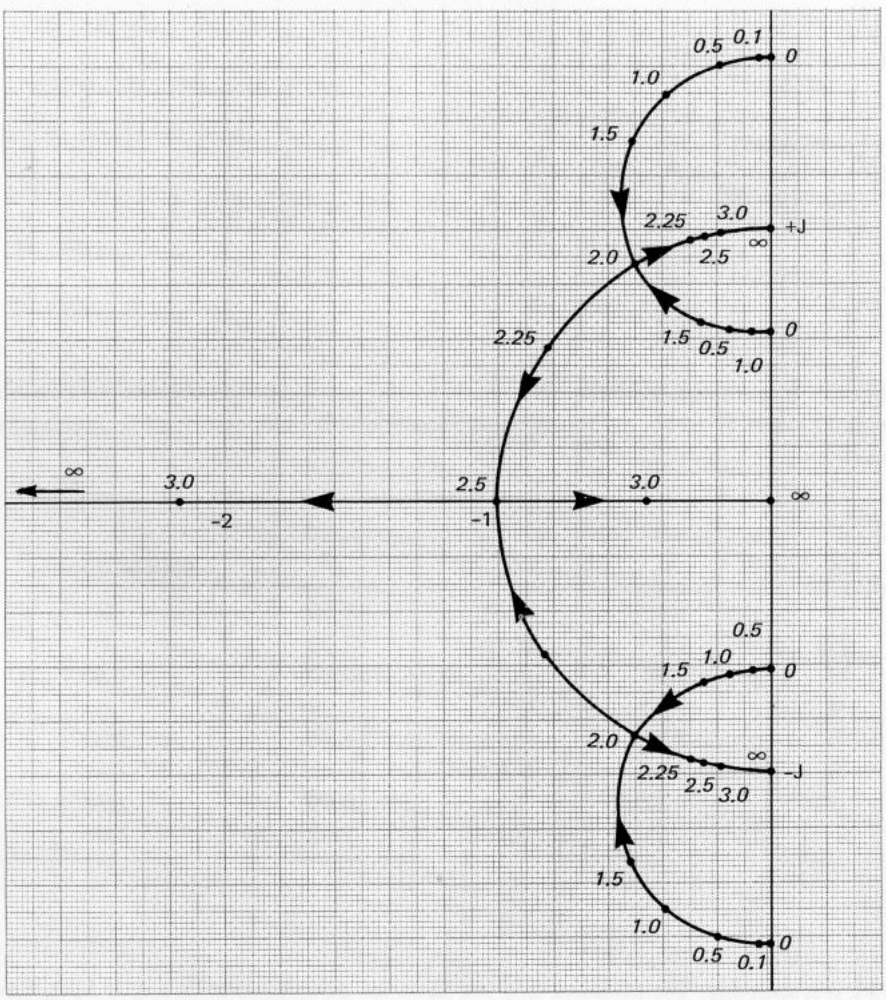

Figure 2 — Pole positions as a function of Q (values of Q in italics) of $1/(s^4 + Qs^3 + Rs^2 + Qs + 1)$ for R = 3

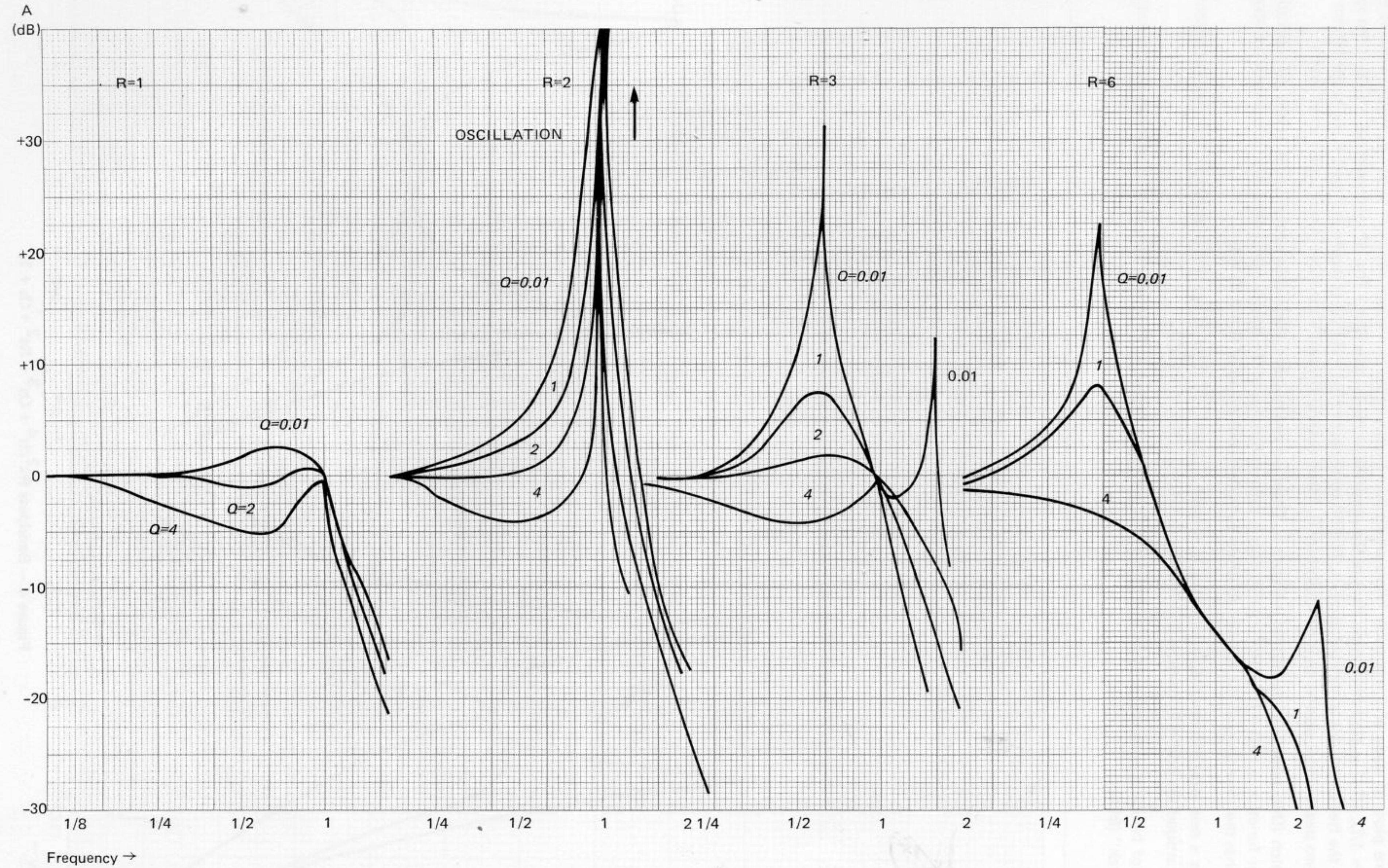

Figure 3 — Low-Pass $1/(s^4 + Qs^3 + Rs^2 + Qs + 1)$

Now to examine plots of some of the possible filters which may be generated for various values of Q and R. Figure 3 shows the LP response $H(s) = 1/D$. (R=2 is plotted anyway despite the presence of the on-axis pole.) The HP response $H(s) = s^4/D$ is the same as the LP except the frequency axis is inverted about 1. The BP response of $H(s) = Rs^2/D$ is shown in Figure 4. As in the LP, for small Q's it possess two peaks whose separation is controlled by R. These doubly resonant filters can give a vocal sort of sound.

The double notch (DN) responses are plotted in Figure 5. There are two notches, the widths controlled by Q and the separation by R. The transfer function is $H(s) = (s^4 + Rs^2 + 1)/D$. This is a generalization of the double notch obtained from a standard four-pole phasor, which has the same form of the transfer function but with fixed values of R = 6, Q = 4.

Figure 6 shows a new kind of response, which may be termed a "peak" (P) filter. It has unit response at high and low frequencies, but two peaks around the center frequency. The transfer function is $H(s) = (1+s^4)/D$.

Another kind of bandpass is available by $H(s) = (s + s^3)/D$, shown in Figure 7. This has a zero at s = j, so it can be designated as a "bandpass/notch" (BPN).

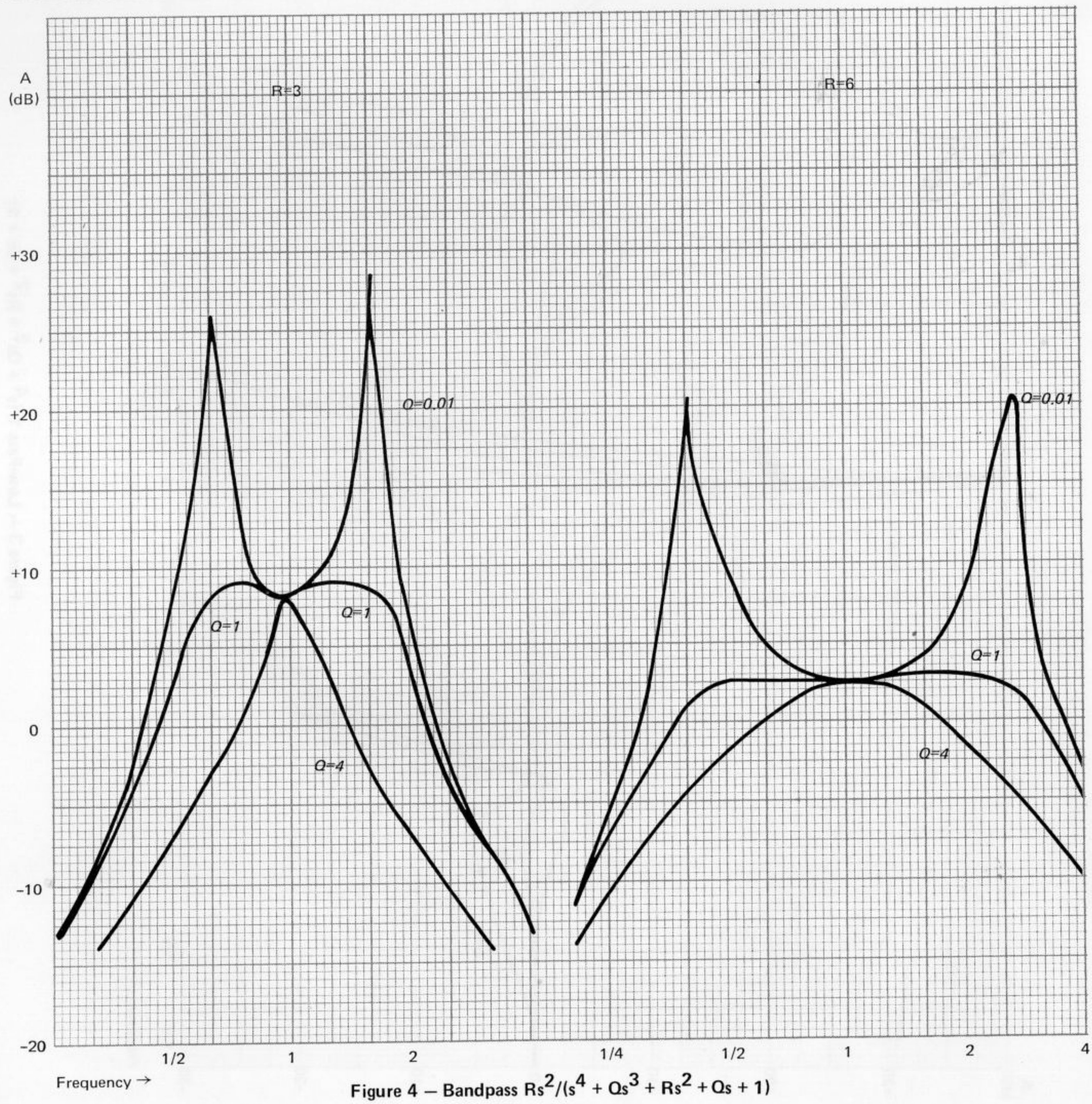

Figure 4 — Bandpass $Rs^2/(s^4 + Qs^3 + Rs^2 + Qs + 1)$

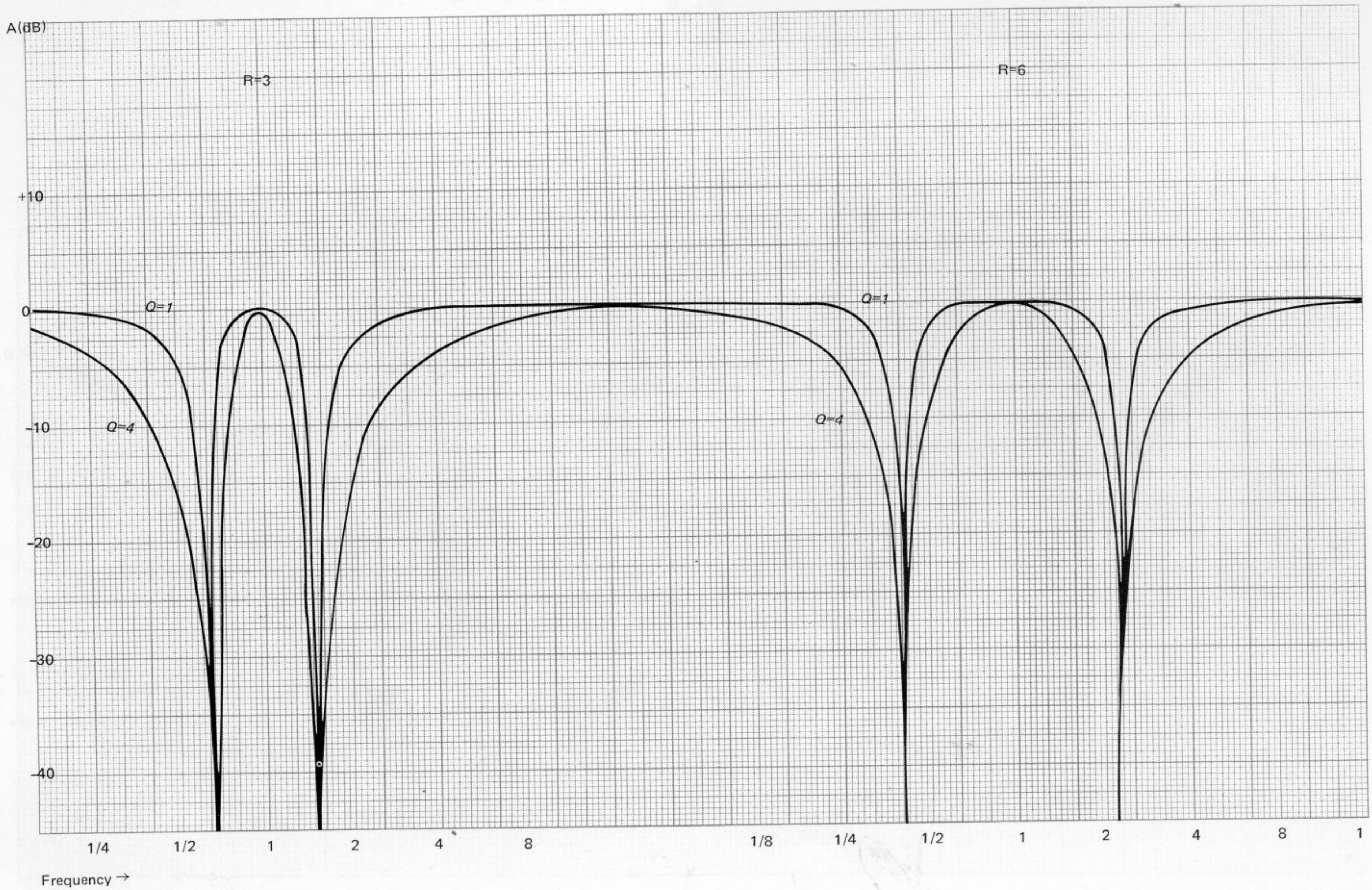

Figure 5 — Double Notch. $(s^4 + Rs^2 + 1)/(S^4 + Qs^3 + Rs^2 + Qs + 1)$

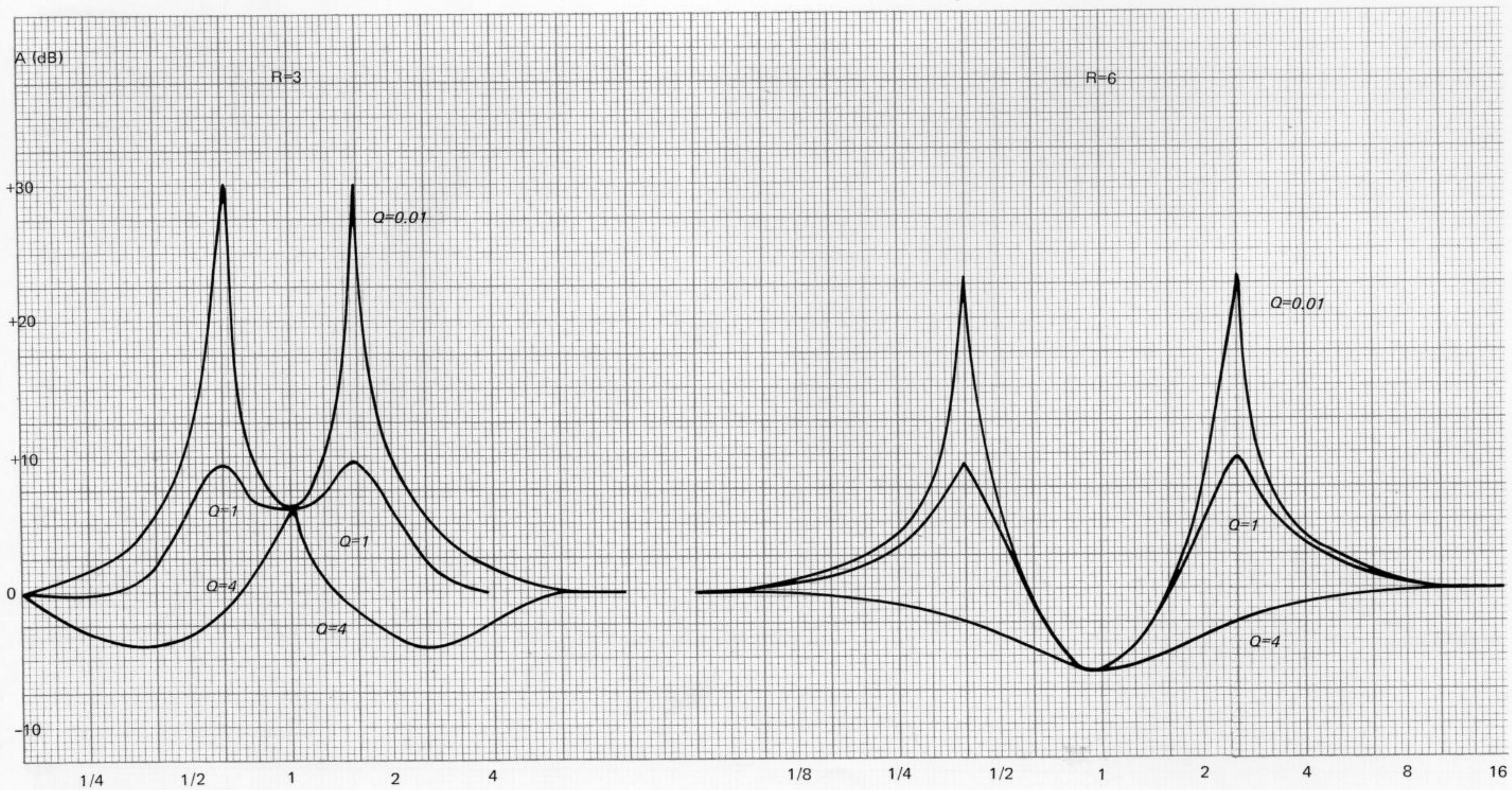

Figure 6 — Peak $\frac{(s^4 + 1)}{(s^4 + Qs^3 + Rs^2 + Qs + 1)}$

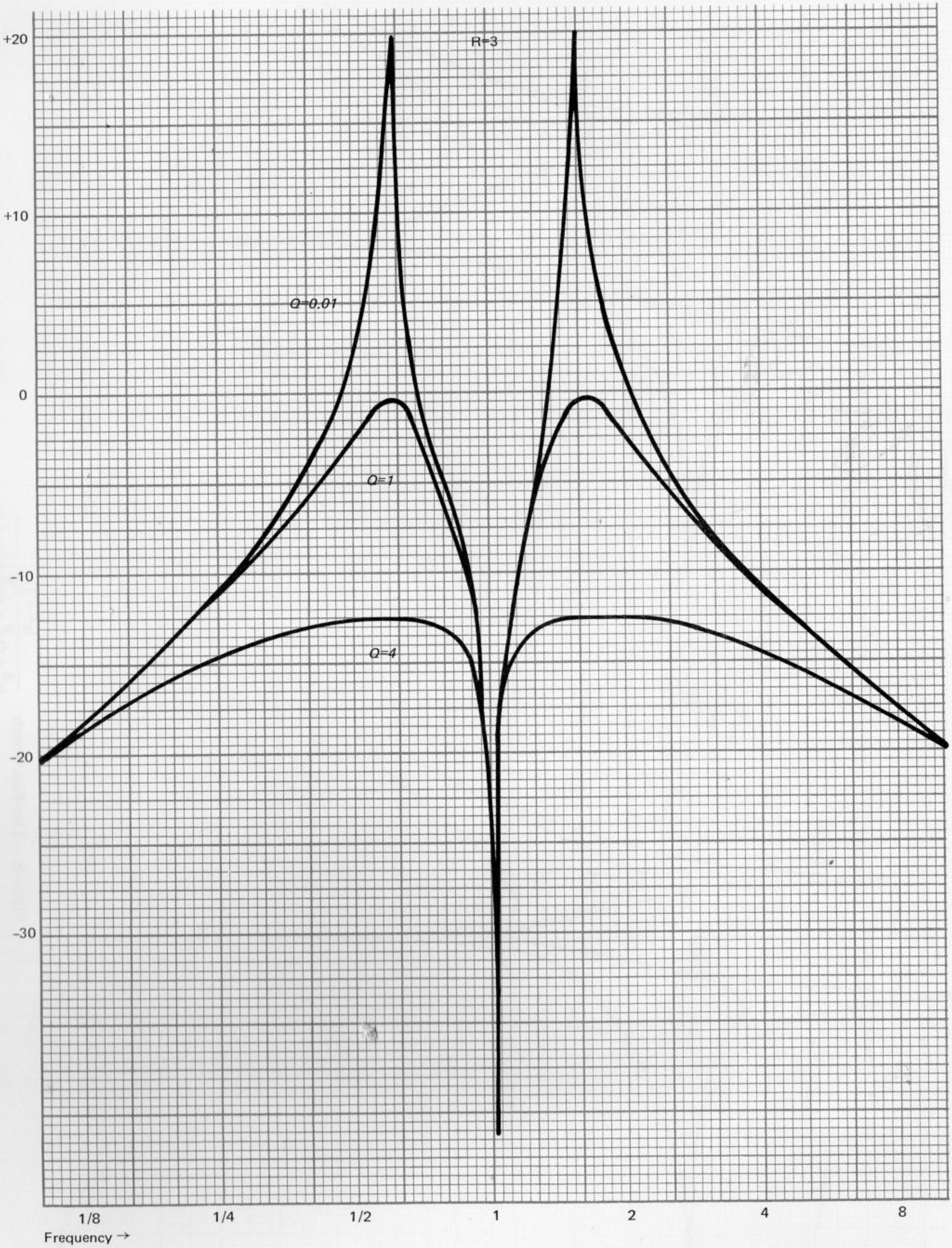

Figure 7 — Bandpass/Notch $\dfrac{(s + s^3)}{(s^4 + Qs^3 + Rs^2 + Qs + 1)}$

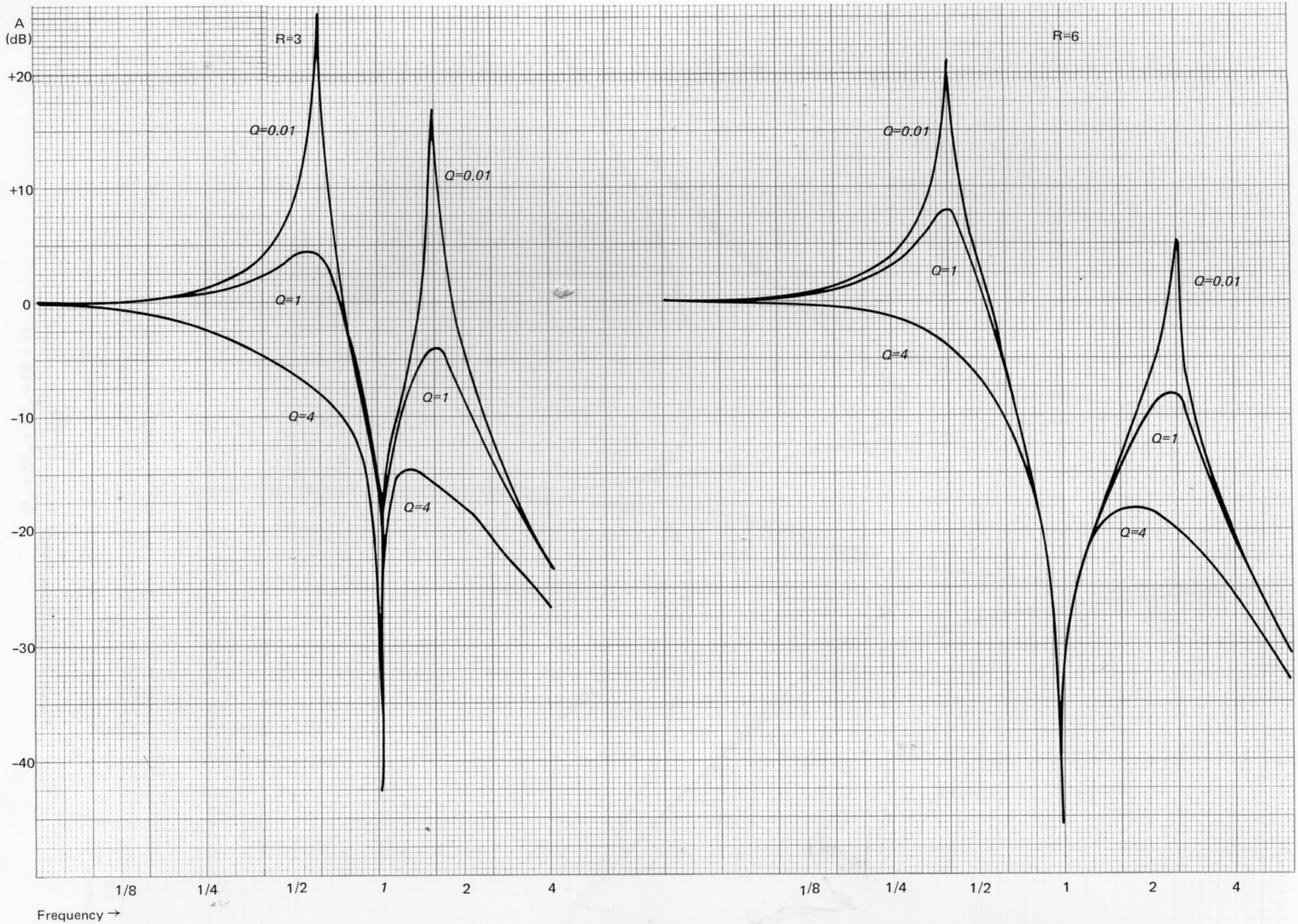

Figure 8 — Low-Pass/Notch $\dfrac{(s^2+1)}{(s^4+Qs^3+Rs^2+Qs+1)}$

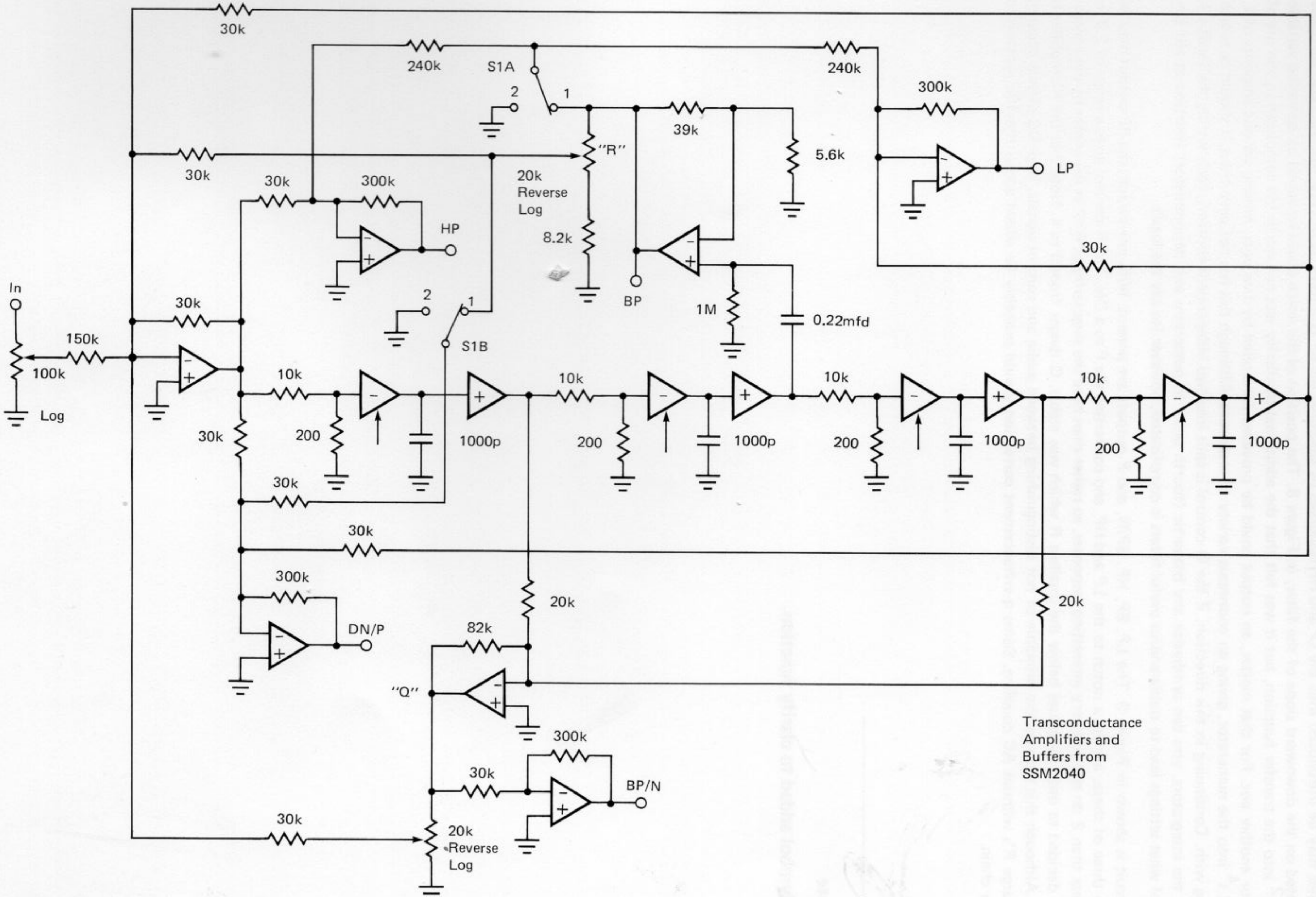

Figure 9 – Fourth-Order State-Variable Filter

Finally, notches may be introduced into the LP and HP responses by summing the BP (s^2) term, as in $(1+s^2)/D$. This also enhances the peak located on the downward slope of the filter, in Figure 8. The position of the notch could be varied by summing variable amounts of s^2 into the transfer function, but it was felt that the additional flexibility was not worth the complication and use of panel space for another pot. For that matter, an output could be provided controlled by five pots, mixing various amounts of 1, s, s^2, s^3, and s^4 into the numerator, giving an enormous variety of responses. Although this has not been tried, it should be worth experimenting with. Continuing in this direction, if the Q control is split into two independent controls, each setting the feedback from one of the integrators, you can synthesize any biquartic (fourth order in numerator and denominator) function at all! Determination of what settings lead to stable and/or useful filters is complicated, however (to say the least).

An actual circuit is shown in Figure 9. The LP, BP, HP, BP/N, and P outputs are present. With one switch, the BP output may be summed into three of these, adding a notch to the LP and HP, and converting the P to a DN. The R control gives a range of 2.3 to 8. R values less than 2 do not give very interesting responses, so rather than have the singularity at R=2 in the center of the control range, it was decided to omit all values below the smallest R which was stable. Q ranges from 0 to 4. Note that the R amplifier is AC coupled. Although this violates the principle of not distinguishing between audio and control signals, large DC offsets occured when using large R's without AC coupling. Some q-enhancement correction should probably be added across the 10k resistors in the integrator chain.

Editor's Note

†Summer Symbol added to clarify function.

Solid State Micro Technology for Music

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I.C. EVALUATION BOARD

Solid State Microtechnology and Eμ Systems of Santa Cruz, California, have collaborated on developing a synthesizer voice demo board using the SSMT music I.C. line. The board is supplied with complete documentation, SSMT I.C.'s, polystyrene caps and Tel Labs resistors for the VCO's. The board is available to O.E.M. manufacturers interested in evaluating the parts through SSMT for \$60. Individuals who have an interest in the board should contact Eμ Systems who also offer assembled and tested boards for \$450.

All the inputs and outputs from the modular blocks on the board are pinned out at the edge connector so that each block can be evaluated individually and interconnected in any manner for system prototyping. Depending on the user's application, some components may be omitted and some inputs ignored. For example the on-board regulator may be deleted if out-board power supplies are available.

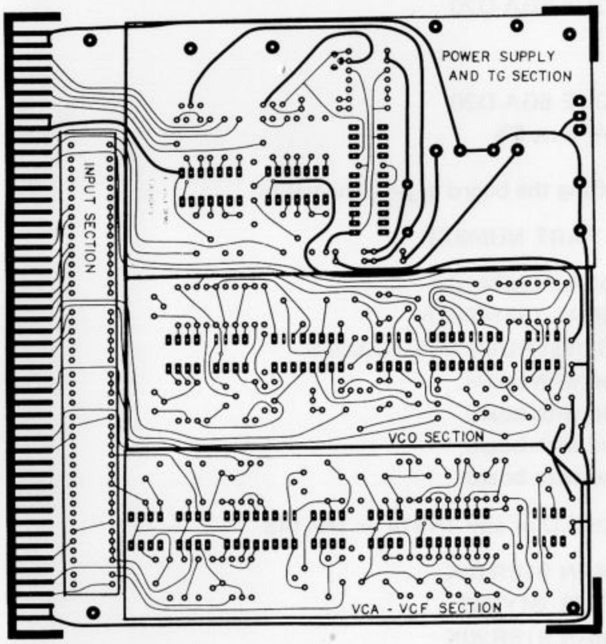
The enclosed documentation includes a complete parts list for the voice demo board, including a manufacturer's part number for each item. Except those items supplied by SSM, most items can be substituted readily with only slight effect on circuit performance. In particular:

The 1% resistors can be substituted for near-valued 5% units with some loss in temperature stability.

The TL082 op-amps can be substituted by 1458's with some loss of bandwidth and oscillator stability with temperature.

The trimmers specified can be freely substituted, but cermet varieties will maintain best tempco.

The typical front panel schematic is a suggestion only, to show how the board can be connected. Infinite variations exist.



Board Layout

Solid State Micro Technology for Music, 2076B Walsh Avenue, Santa Clara, CA 95050, USA
(408) 248-0917 Telex **171189**

PARTS LIST

QTY	PART	MFR'S PART NUMBER
1	Circuit Board	Available from E-mu — $50 by itself
1	Edge Conn.	TEKA TP5-W01-50 or any S-100 type.

The following parts are required for the on-board power supply. None are required if +/– 15V are supplied to the board externally. The user must supply fused 110VAC to the board's power connector.

1	Power Conn.	MOLEX 09-65-1031
1	Mating Conn.	MOLEX 09-50-3031 with 3 08-50-0106 pins
1	Transformer	SIGNAL TRANSFORMER ST 4-36
2	200uF 40V Cap	SPRAGUE 500D208G050FF7
4	1N4002 Diode	MOTOROLA
2	2N4923	MOTOROLA
2	Heatsinks	THERMALLOY 6073
2	Hdwe for above	2 ea. 4-40x1/2" screw, nut, lockwasher
1	741 opamp	NATIONAL LM741CN
1	723 regulator	NATIONAL LM723CN
1	2N3904	MOTOROLA
4	3.01K 1% Res	DALE RN55D
1	3.32K 1% Res	DALE RN55D
1	9.09K 1% Res	DALE RN55D
2	3.9 ohm Res	Any 1/4 watt 5%
1	200 ohm	Any 1/4 watt 5%
1	470 pF Cap	SPRAGUE 5GA-T47

The following parts are required depending on filter option:

LOWPASS:

4	1000 pF Cap	SPRAGUE 5GA-D10
5	10K Res	Any 1/4 watt 5%

HIGHPASS:

4	2000 pF Cap	SPRAGUE 5GA-D20

ALLPASS (Phase-shifter):

4	2000 pF Cap	SPRAGUE 5GA-D20
5	10K Res	Any 1/4 watt 5%

The remaining parts are required for stuffing the board as per schematics:

QTY	PART	MFR'S PART NUMBER
1	741 opamp	NATIONAL LM741CN
4	1458 dl opamp	NATIONAL LM1458CN
4	TL082 dl opamp	TEXAS INSTRUMENTS TL082CP
1	SSM2020	Supplied with board.
2	SSM2030	Supplied with board.
1	SSM2040	Supplied with board.
2	SSM2050	Supplied with board.
8	1N914 Diode	FAIRCHILD or any 1N914 or 1N4148
2	10K Trimmer	BECKMAN 91BR10K
2	20K Trimmer	SPECTROL 64Y203
1	20K Trimmer	BECKMAN 91BR20K
4	100K Trimmer	BECKMAN 91BR100K
2	121 ohm 1% R	DALE RN55D
2	54.9K 1% Res	DALE RN55D
2	90.9K 1% Res	DALE RN55D

PARTS LIST (continued)

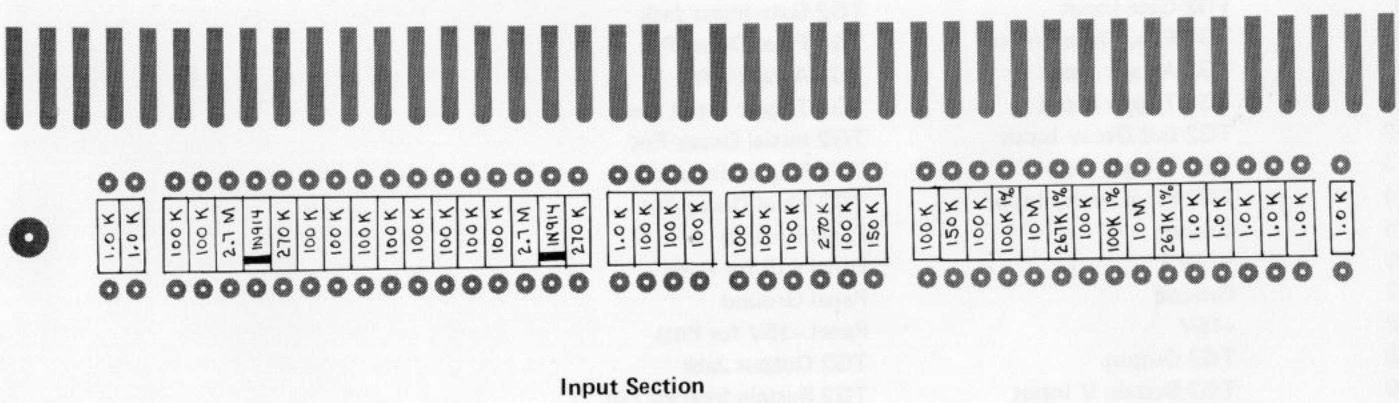
QTY	PART	MFR'S PART NUMBER
2	100K 1% Res	DALE RN55D
2	267K 1% Res	DALE RN55D
3	1.0K Tempco R	TEL LABS Q-81 Supplied with board.
3	5 pF Cap	SPRAGUE 10TSV33
3	22 pF Cap	SPRAGUE 10TSQ22
4	100 pF Cap	SPRAGUE 10TST10
4	1000 pF Cap	SPRAGUE 5GAD10
2	1000 pF PolyS	MALLORY SXM110 Supplied with board.
2	0.01 uF Cap	SPRAGUE TG-S10
14	0.1 uF Cap	SPRAGUE 3CZU104D8050C5
2	10 uF Tant Cap	SPRAGUE 196D106X0035PE4
4	200 ohm Res	All following are any 1/4 watt 5%
11	1.0K	
6	2.2K	
2	7.5K	
10	10K	
4	15K	
4	20K	
4	22K	
2	27K	
8	47K	
1	56K	
1	91K	
29	100K	
2	150K	
5	270K	
6	330K	
2	470K	
2	1.5M	
2	2.2M	
2	2.7M	
2	10M	

PINOUTS

PIN #	SIGNAL NAME	TYPICAL CONNECTION
1-4	Ground	Panel Ground
5	TG1 Attack Input	TG1 Attack Pot
6	TF1 Gate Input	TG1 Gate Input Jack
7	TG1 Init Decay Input	TG1 Initial Decay Pot
8	TG2 Gate Input	TG2 Gate Input Jack
9	TG1 Final Decay Input	TG1 Final Decay Pot
10	TG2 Attack Input	TG2 Attack Pot
11	TG1 Trigger Input	TG1 Trigger Input Jack
12	TG2 Init Decay Input	TG2 Initial Decay Pot
13	TG2 Trigger Input	TG2 Trigger Input Jack
14	TG2 Final Decay Input	TG2 Final Decay Pot
15	Ground	Panel Ground
16	+15V	Panel +15 for Pots
17	Ground	Panel Ground
18	-15V	Panel -15V for Pots
19	TG2 Output	TG2 Output Jack
20	TG2 Sustain V Input	TG2 Sustain Voltage Pot

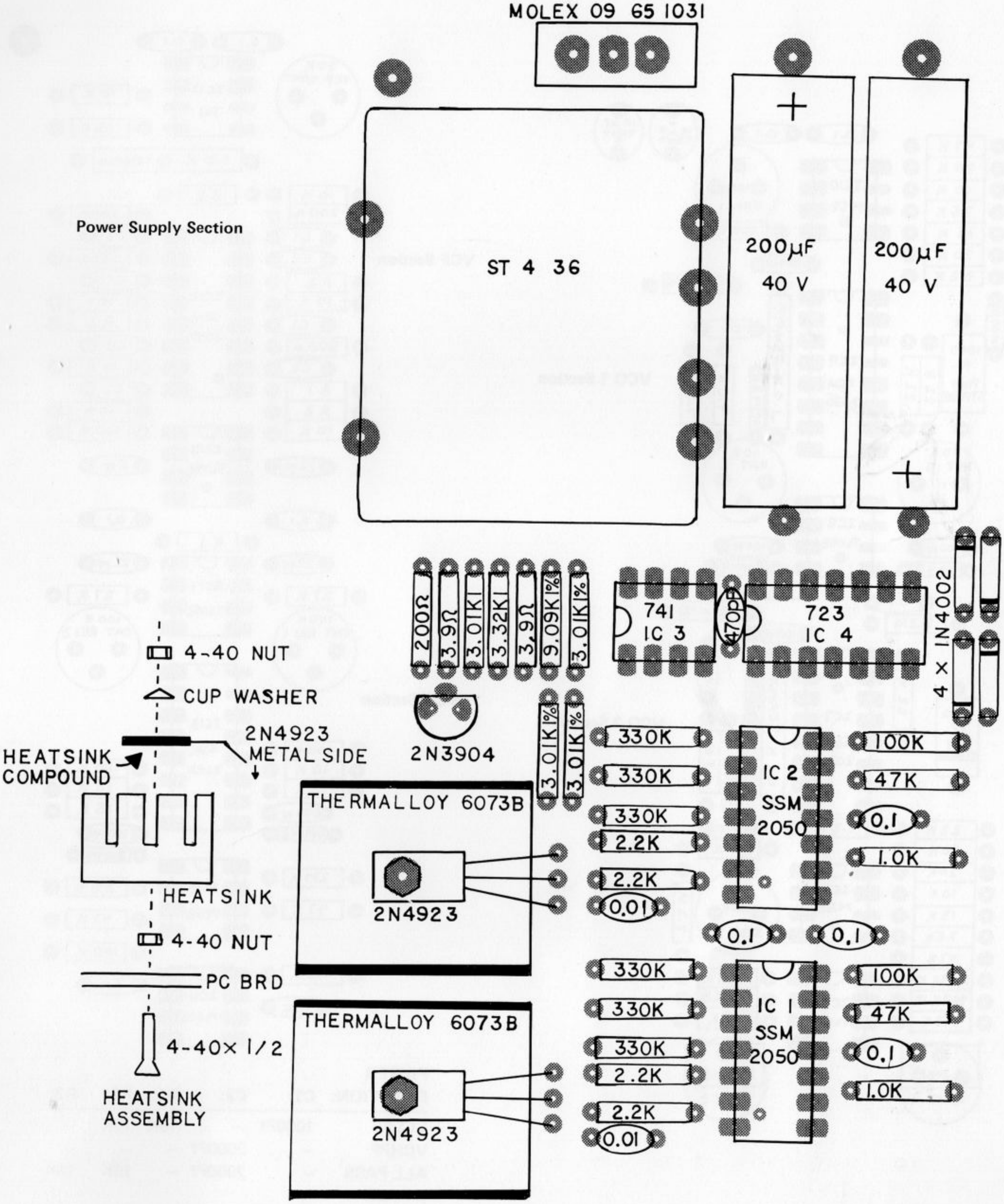
PINOUTS (continued)

PIN #	SIGNAL NAME	TYPICAL CONNECTION
21	TG1 Output	TG1 Output Jack
22	TG1 Sustain V Input	TG1 Sustain Voltage Pot
23	VCO1 Sync Input	VCO1 Sync Input Jack
24	VCO1 Triangle Output	VCO1 Triangle Output Jack
26	VCO1 Sawtooth Output	VCO1 Sawtooth Output Jack
28	VCO1 Pulse Output	VCO1 Pulse Output Jack
30	VCO2 Pulse Output	VCO2 Pulse Output Jack
32	VCO2 Sawtooth Output	VCO2 Sawtooth Output Jack
34	VCO2 Triangle Output	VCO2 Triangle Output Jack
37	VCO1 Init Freq Coarse	VCO1 Initial Freq Coarse Pot
38	VCO1 Keyboard Input	VCO1 Keyboard Input Jack
39	VCO1 Init Freq Fine	VCO1 Initial Freq Fine Pot
40	VCO1 Freq Mod Input	VCO1 Frequency Modulation Input Attenuator
43	VCO2 Init Freq Coarse	VCO2 Initial Freq Coarse Pot
44	VCO2 Keyboard Input	VCO2 Keyboard Input Jack
45	VCO2 Init Freq Fine	VCO2 Initial Freq. Fine Pot
46	VCO2 Freq Mod Input	VCO2 Frequency Modulation Input Attenuator
47	VCO2 Pulse Width Cntl	VCO2 Pulse Width Pot
48	VCO2 Pulse Width Input	VCO2 Pulse Width Mod Input Attenuator
49	VCO2 Linear FM Input	VCO2 Linear FM Input Attenuator
51	VCO1 Linear FM Input	VCO1 Linear FM Input Attenuator
52	VCO2 Sync Input	VCO2 Sync Input Jack
53	VCO1 Pulse Width Cntl	VCO2 Pulse Width Pot
54	VCO1 Pulse Width Input	VCO1 Pulse Width Mod Input Attenuator
55	VCF Initial Freq Input	VCF Initial Frequency Pot
56, 58, 60	VCF Freq Cntl Inputs	VCF Freq Cntl Input Attenuators
61	VCF Resonance Input	VCF Resonance Control
62, 64, 66	VCF Signal Inputs	VCF Signal Input Attenuators
69	VCA1 Exp'l Cntl Input	VCA1 Exp'l Gain Cntl Input Attenuator
70	VCA1 Output	VCA1 Output Jack
71	VCA1 Init Gain Input	VCA1 Initial Gain Pot
72, 74	VCA1 Gain Cntl Inputs	VCA1 Gain Control Input Attenuators
76, 78, 80	VCA1 Signal Inputs	VCA1 Signal Input Attenuators
82, 84, 86	VCA2 Signal Inputs	VCA2 Signal Input Attenuators
89	VCA2 Init Gain Input	VCA2 Initial Gain Pot
90, 92	VCA2 Gain Cntl Inputs	VCA2 Gain Control Input Attenuators
93	VCA2 Exp'l Cntl Input	VCA2 Exp'l Gain Cntl Input Attenuator
94	VCA2 Output	VCA2 Output Jack
95	VCF Resonance Output	VCF Resonance Control
96	VCF Output	VCF Output Jack
97-100	Ground	Panel Ground

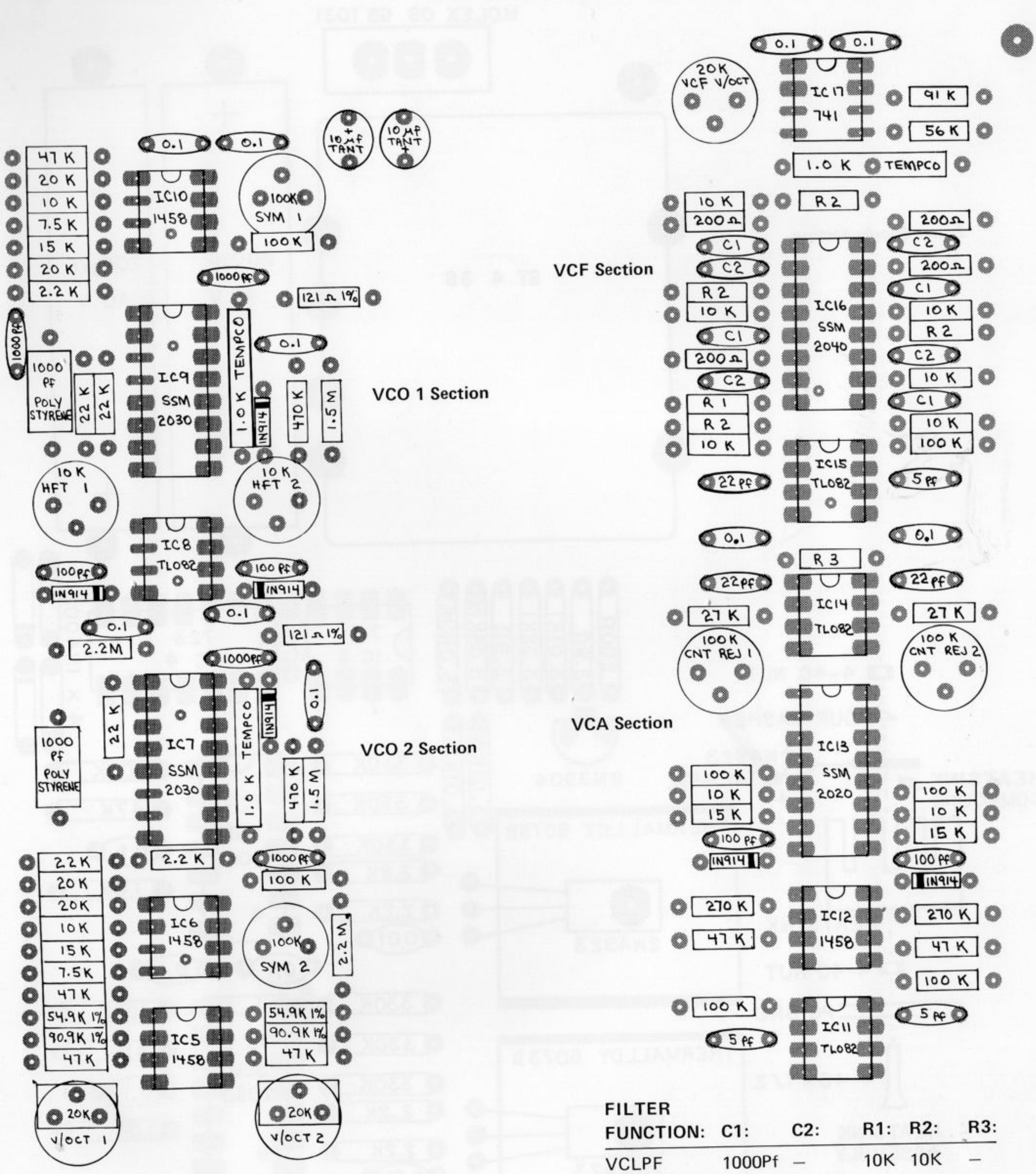


Input Section

Power Supply Section

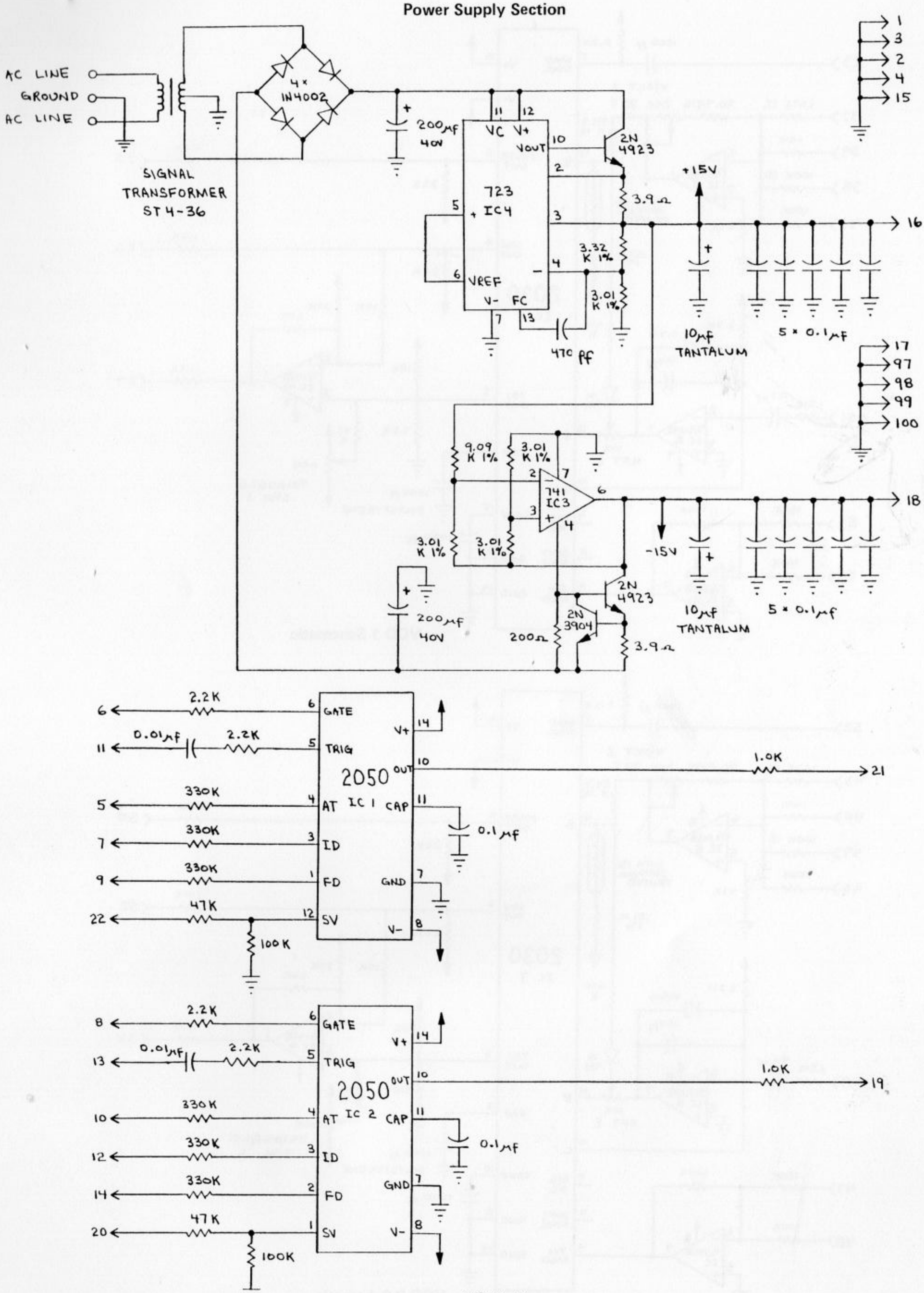


TG Section

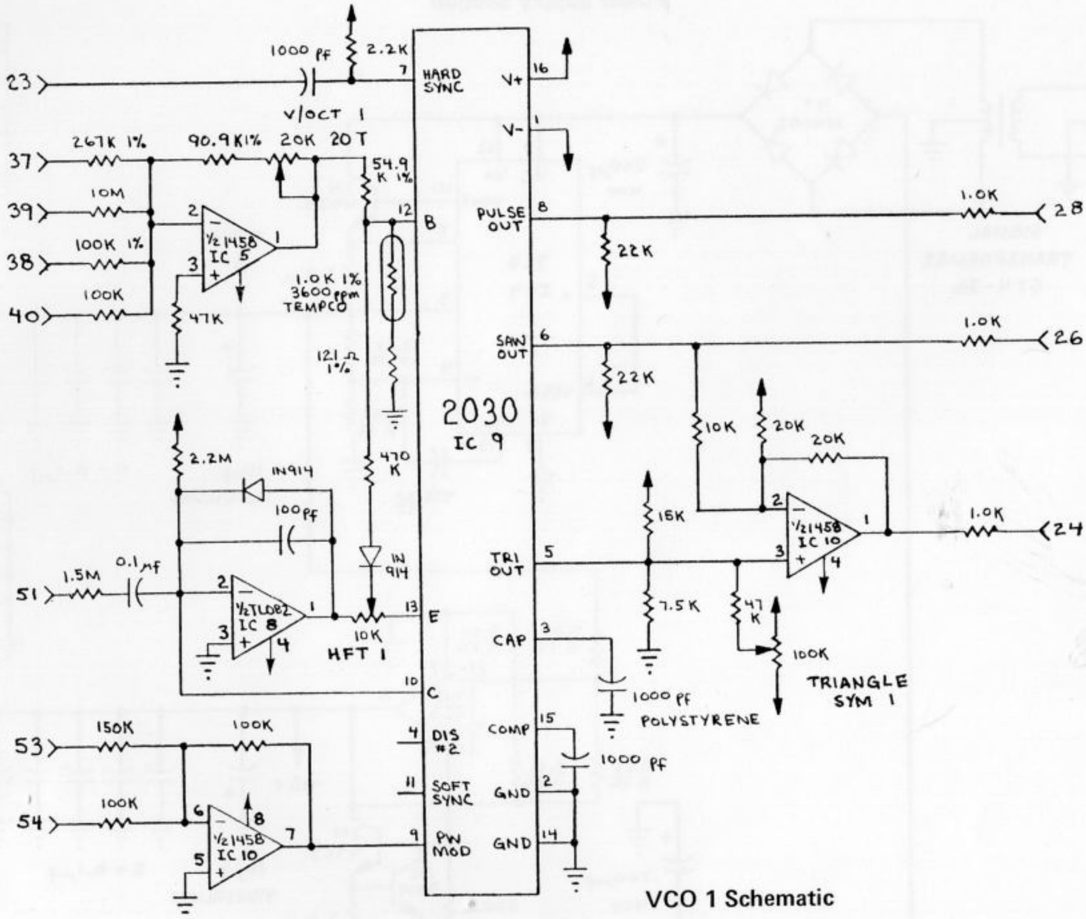


FILTER FUNCTION:	C1:	C2:	R1:	R2:	R3:
VCLPF	1000Pf	–	10K	10K	–
VCHPF	–	2000Pf	–	–	–
ALL PASS (PHASE SHIFT)	–	2000Pf	–	10K	10K

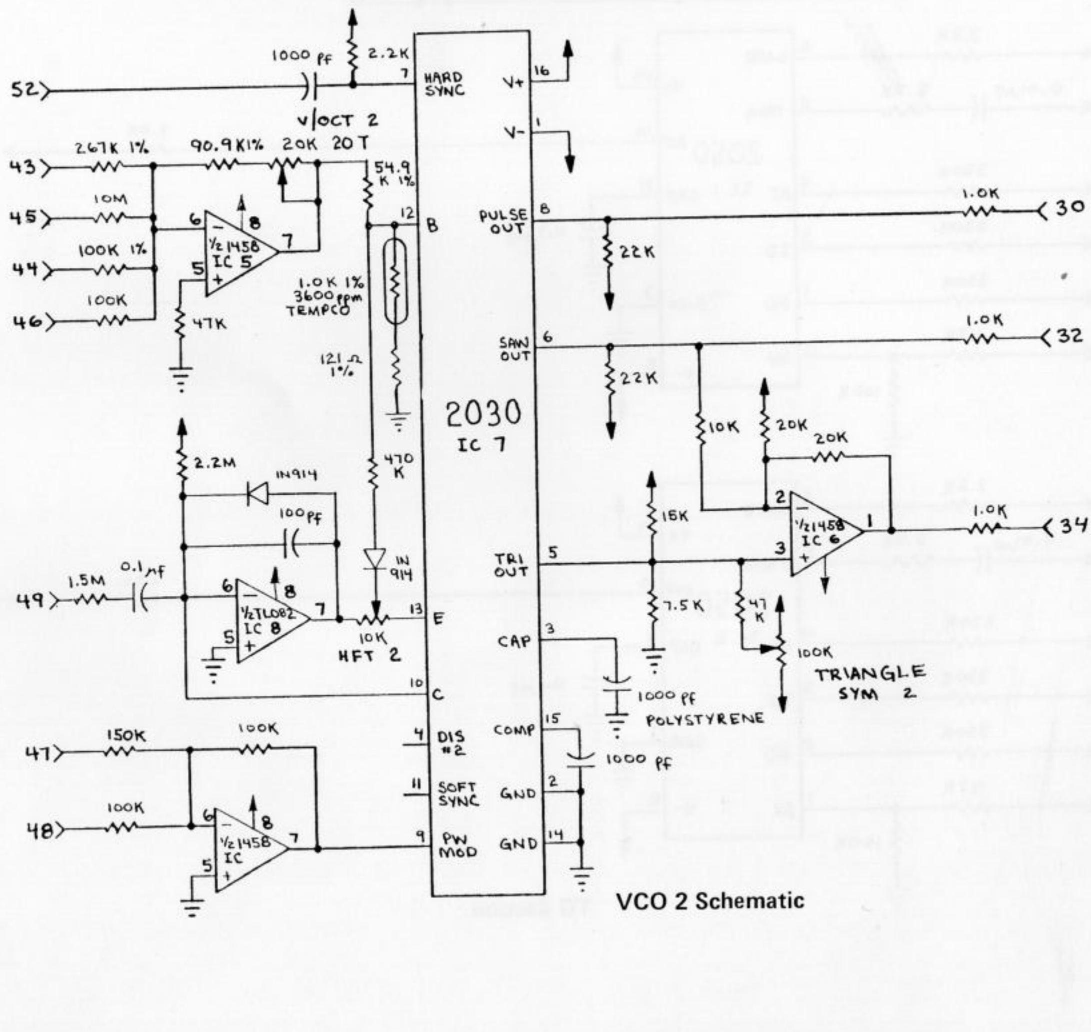
Power Supply Section



TG Section

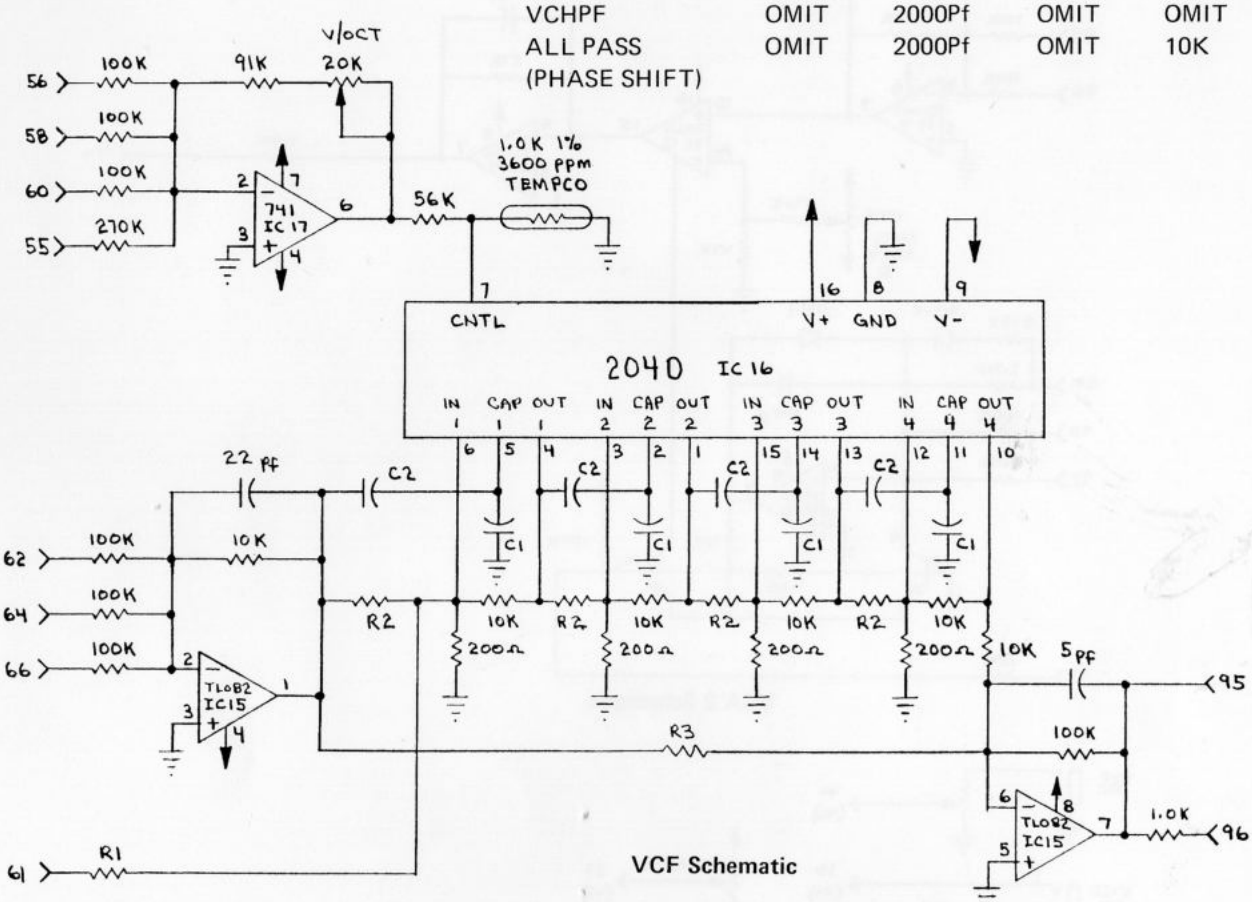


VCO 1 Schematic

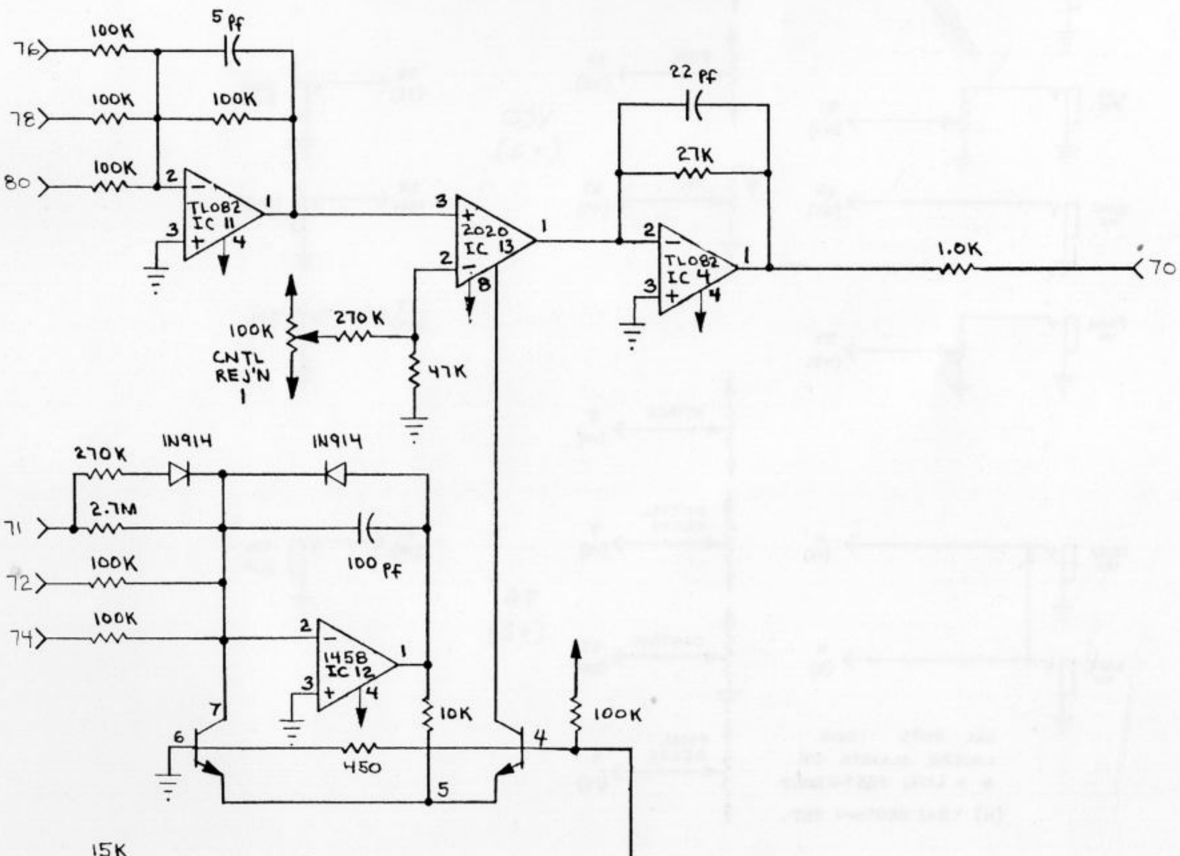


VCO 2 Schematic

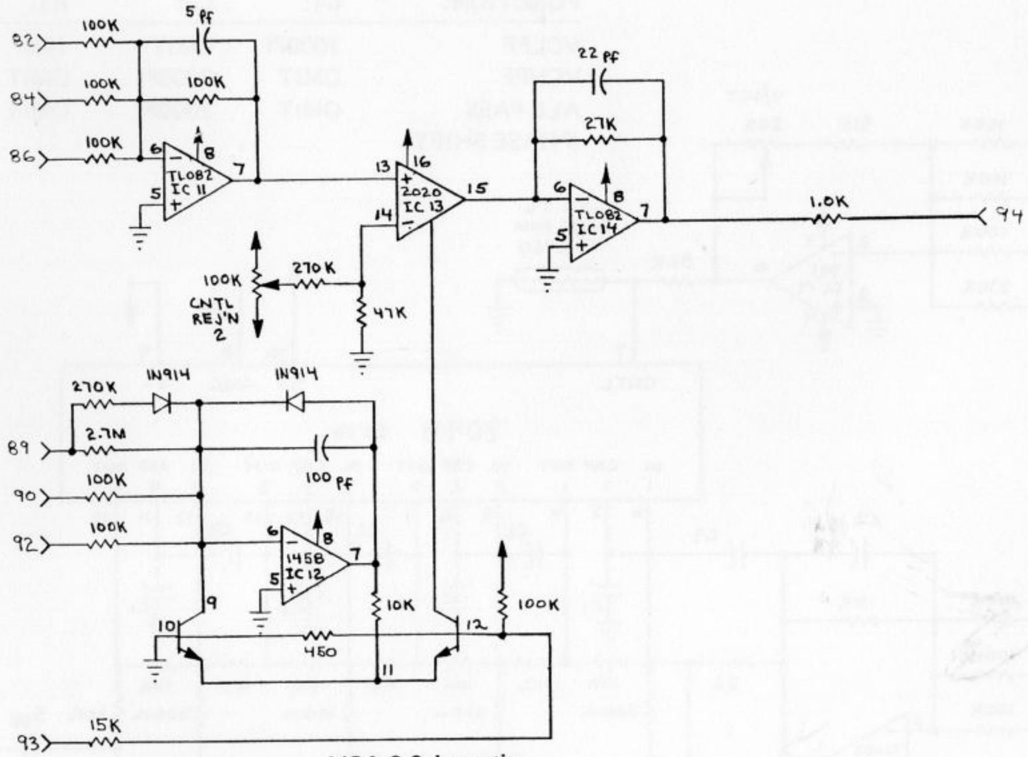
FUNCTION:	C1:	C2:	R1:	R2:	R3:
VCLPF	1000Pf	OMIT	10K	10K	OMIT
VCHPF	OMIT	2000Pf	OMIT	OMIT	OMIT
ALL PASS (PHASE SHIFT)	OMIT	2000Pf	OMIT	10K	10K



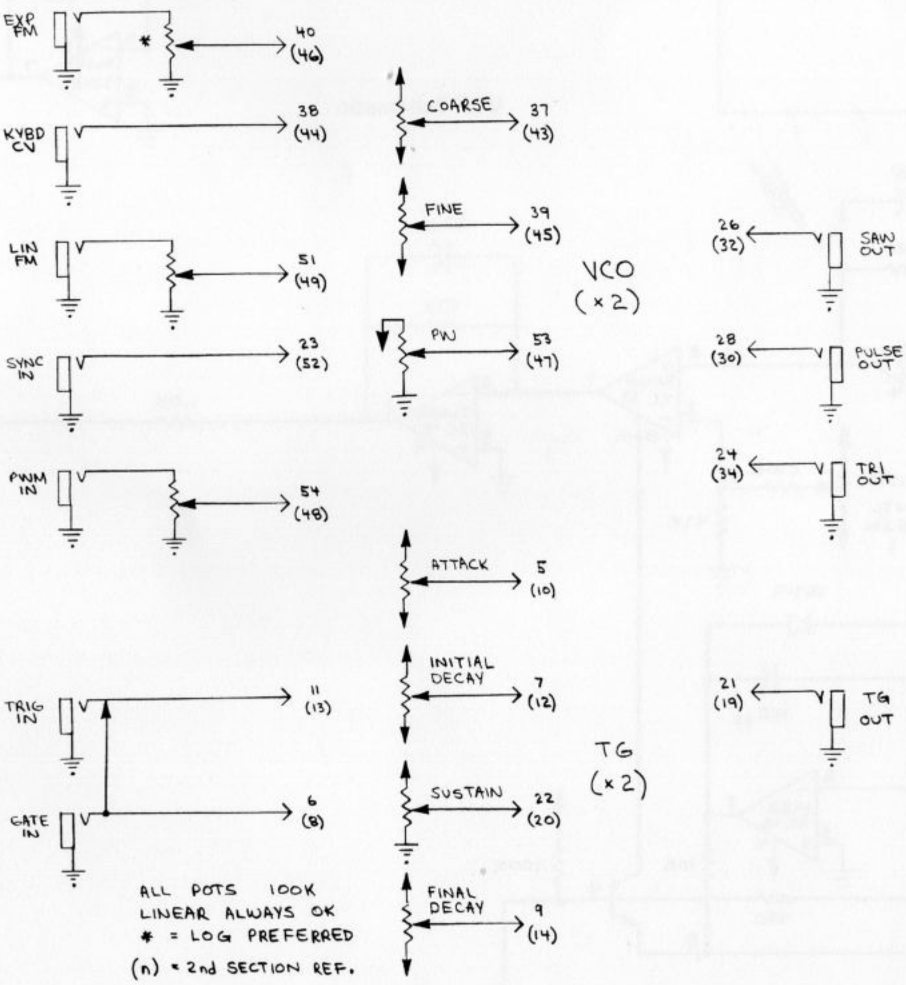
VCF Schematic



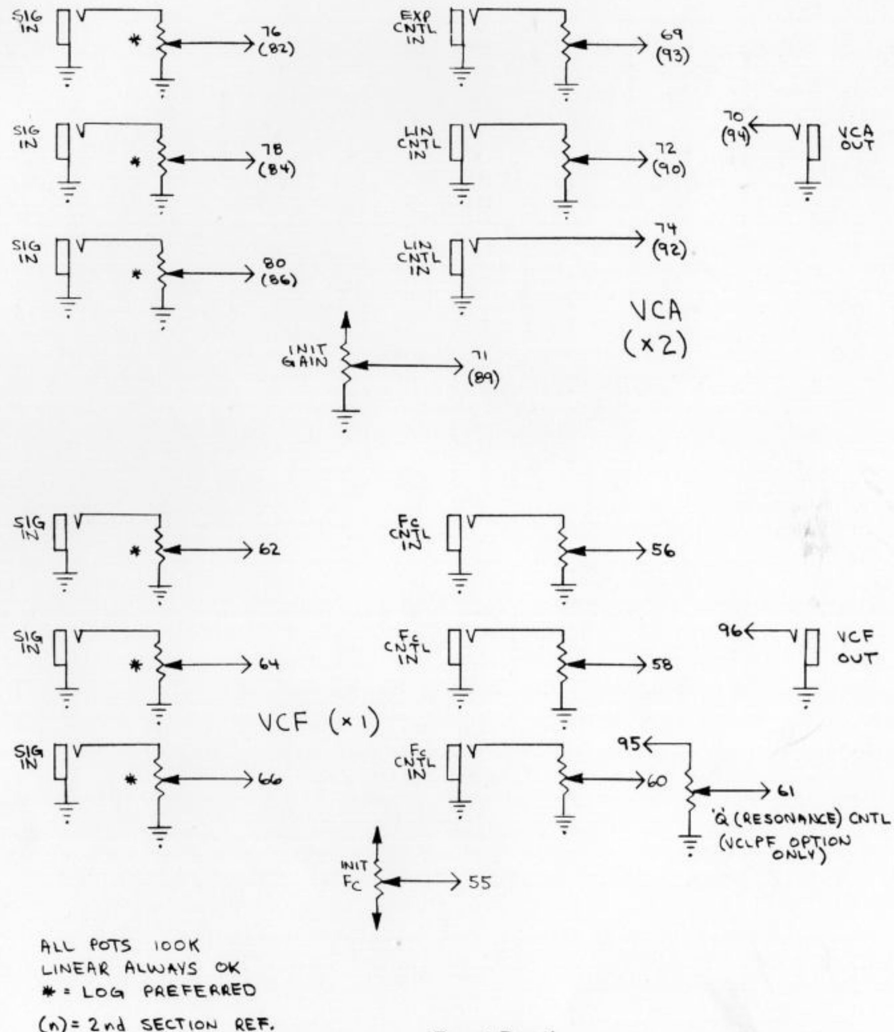
VCA 1 Schematic



VCA 2 Schematic



Panel Controls



Front Panel

TRIM PROCEDURE

OSCILLATOR TRIMS (2 each):

SYM: Monitor triangle output of VCO with 'scope or ear. With 'scope, adjust for minimal glitch on triangle waveform. With ear, adjust for minimal buzz; this is best done at lower frequencies.

V/OCT: Monitor sawtooth output of VCO. Adjust for precisely 1 octave shift from 100 to 200 Hz for a 1 volt change at the keyboard input. This is usually best done by ear, but a good 'scope can be used. The 1 volt source is usually the system keyboard if one is used, but any precise 1 volt reference will do. The most important thing is that the oscillators track.

HFT: Do this after V/OCT. This trim optimizes the VCO tracking at high frequencies. While monitoring the sawtooth output of the VCO, adjust for precisely 1 octave shift from 2000 to 4000 Hz for a 1 volt change at the keyboard input. You may wish to re-trim the V/OCT trim after you first adjust the HFT.

FILTER:

V/OCT: With the lowpass configuration, turn up the "Q" until the filter oscillates. Adjust so that a 1 volt change on a frequency control input results in a 1 octave shift in frequency from 250 to 500 Hz. With high pass or allpass configuration the trim is much more difficult. It is usually satisfactory to just center the trimmer, but you may either adjust the trimmer for precisely 18.02 mV change at pin 7 of IC 16 at room temperature for 1 volt change on a frequency control input, or adjust such that the output waveform from a VCO tracking the filter stays constant.

VCA (2 each):

CNT R: Connect a 1 KHz triangle waveform to a gain control input, and monitor the output of the VCA. Adjust for minimal audio output. The signal inputs to the VCA must be open.

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